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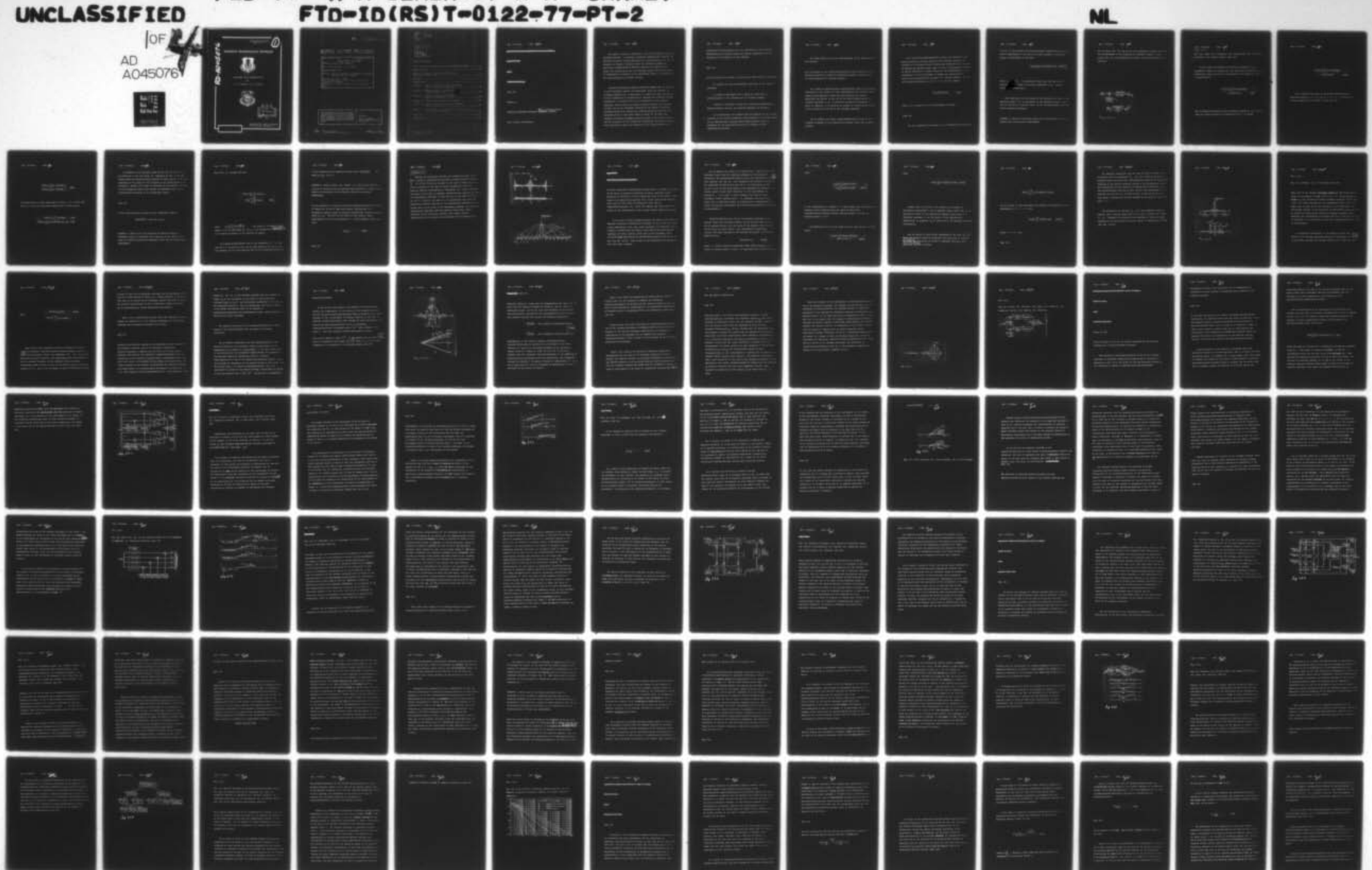

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PART 2 O

FOREIGN TECHNOLOGY D

BROADBAND RADIO COMMUNICATI

by

A. M. Semenov, A. A. Sikar

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BROADBAND RADIO COMMUNICATION

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Chapter 3.

MUTUALLY CORRELATED BROADBAND ~~RADIOLINK~~ RADIO COMMUNICATION SYSTEMS.

§3.1. General information.

The present chapter is dedicated to one of the possible forms of broadband systems - to mutually correlated communicating systems. The distinctive special feature/peculiarity of such systems is the use in receptor for making a decision about the transmitted symbol of short-term crosscorrelation function between the adopted sum of signal and interferences and the reference signal. The latter either is regenerated in receptor or the information about it is embedded in the characteristics of matched filters.

Mutually correlated broadband radiolink systems make it possible to most completely realize the advantages, which are inherent in broadband systems and, in particular, to ensure the high correctness of the transmission of information in channels with multiple-pronged propagation. Is achieved this by the separation of the incoming ray/beams, during which are removed selective fadings and phenomena of echo, and by the subsequent addition of the energy of separate ray/beams taking into account the characteristics of multiple-beam characteristics at each given moment of time. At the same time mutually correlated broadband systems provide to the certain degree and the reticence of the transmitted information, the higher freedom from interference under the influence of the various kinds of the

concentrated interferences, etc. The realization of the indicated advantages can be reached during the correct construction of the diagrams of the receptors of such systems.

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For this provision necessary to know how to solve series of problems:

- to isolate one and, consequently, also each of the incoming ray/beams;

- to supply requirements and to carry out them with the synchronization of the reference and of received signals;

- correct to construct, taking into account the parameters of multiple-pronged channel, the decisive schematic of receptor;

- to rate/estimate the freedom from interference of the diverse variants of the decisive diagrams, which consider to a certain degree of the characteristic multiple beam characteristics, which is necessary for the reasonable selection of receptor in each communication channel.

All these questions are the object/subject of the examination of this chapter.

§3.2. Principle of the isolation/liberation of one ray/beam and the decisive diagrams of single-ray reception/procedure. Conditions of the synchronization of received signals.

As a result of multiple-beam characteristics under the practical conditions of the conduct of radio communication for receptor affect several electromagnetic vibrations with different and random amplitudes and phases. To get rid of the interference between the incoming ray/beams, i.e., of selective fadings and phenomenon of echo, is possible, if we achieve the purpose of isolation/liberation in the receptor only of one of the incoming ray/beams.

Let us examine, as occurs isolation/liberation of one of the incoming ray/beams in the communicating system, which uses broadband signals.

As it was established/installed in §2.7, the necessary stage during processing received signal in the decisive schematic of receptor is its passage either through the correlator (Fig. 3.2.1a), or through the matched filter (Fig. 3.2.1b). Instantaneous the value of output potentials of these devices at the torque/moment of the termination of the cell/element of signal are proportional to short-term crosscorrelation function X_r between adopted x (t) and the supporting/reference $z_r(t)$, r = 1; 2, by the signals:

$$X_r = \frac{1}{T}\int_0^T x(t)\,z_r(t)\,dt, \qquad (3.2.1)$$

where T is a duration of the cell/element of signal.

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Let was transmitted the signal $z_r(t)$ determined by expression

(1.1.1). In the presence of multiple-pronged propagation and with the known torque/moment of the arrival of each ray/beam the received signal is represented in the form

$$x(t)=\sum_{i=1}^{n}\mu_i z_r(t-\Delta t_i)+\xi(t), 0\leqslant t<T, \quad (3.2.2)$$

where μ_i and Δt_i - transmission factor and time lag in i-th ray/bea[illegible] is a number of incoming ray/beams; ξ (t) - normal fluct[illegible] interference.

After admission by the diagram of the correlator of Fig. 3.2.1a received signal x (t) is multiplied by the reference signal, equal with an accuracy to constant factor $z_r(t)$ and synchronized with one of the incoming ray/beams [1].

FOOTNOTE [1]. Here and throughout there are in form binary (r = 1; 2) systems with active pause. ENDFOOTNOTE.

Let us assume that the time lag on each ray/beam is counted off from the torque/moment of the beginning of reference signal, so that values Δt_i in relationship/ratio (3.2.2) can be both positive and negative.

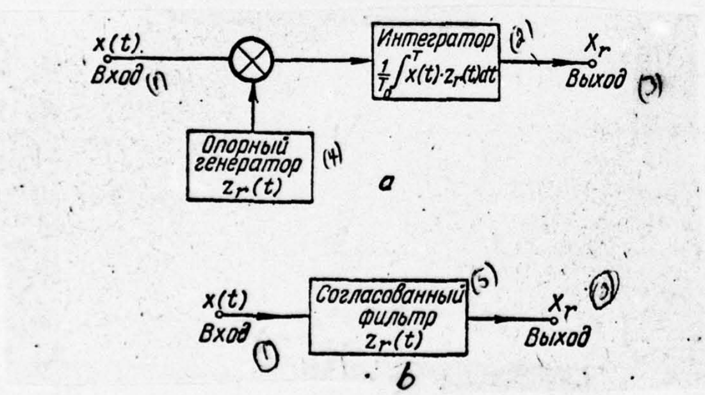

Fig. 3.2.1.

Fig. 3.2.1.

Key: (1). Input. (2). Integrator. (3). Output/yield. (4). Reference oscillator. (5). Matched filter. Page 129.

If the signal of reference oscillator corresponds to the transmitted signal (is agreed with it), then after multiplication and integration output potential of correlator at the moment of time t = T proportionally

$$X_r = \sum_{i=1}^{n} \mu_i \left[\frac{1}{T} \int_0^T z_r(t - \Delta t_i)\, z_r(t)\, dt \right] + + \frac{1}{T} \int_0^T \xi(t)\, z_r(t)\, dt. \quad (3.2.3)$$

But if supporting/reference and transmitted signals are not agreed, then the output potential of correlator with t = T equally

$$X_l = \sum_{i=1}^{n} \mu_{i} \left[\frac{1}{T} \int_0^T z_r(t - \Delta t_i)\, z_l(t)\, dt \right] + \\ + \frac{1}{T} \int_0^T \xi(t)\, z_l(t)\, dt. \tag{3.2.4}$$

Let us examine the terms of short-term crosscorrelation functions between $z_r(t-\Delta t_i)$ and $z_r(t)$, $z_l(t)$ in (3.2.3) and (3.2.4) depending on the value of time lag Δt_i:

$$\left.\begin{aligned} R_l(\Delta t_l) &= \frac{1}{T}\int_0^T z_r(t-\Delta t_l)\, z_r(t)\, dt; \\ R_l'(\Delta t_l) &= \frac{1}{T}\int_0^T z_r(t-\Delta t_l)\, z_l(t)\, dt. \end{aligned}\right\} \qquad (3.2.5)$$

By substituting in these expressions of value $z_r(t)$ and $z_r(t-\Delta t_l)$ from (1.1.1) and by fulfilling integration, we will obtain:

$$R_l(\Delta t_l) = \frac{1}{2}\sum_{k=k_1}^{k_2} A_{rk}^2 \cos k\omega_0 \Delta t_l; \qquad (3.2.6)$$

$$R_l'(\Delta t_l) = \frac{1}{2}\sum_{k=k_1}^{k_2} A_{rk} A_{lk} \cos (k\omega_0 \Delta t_l + \varphi_{rk} - \varphi_{lk}). \qquad (3.2.7)$$

We examine in the beginning terms of the type $R_i(\Delta t_i)$ It is not difficult to see that value R_i depending on Δt_i in no way differs from the autocorrelation function of form (2.5.7), i.e., the dependence of the power of i-th ray/beam at the output/yield of the correlator, matched with signal as function of the relative time lag of the transmitted signal with respect to reference signal is the autocorrelation function of the transmitted signal.

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At the uniform spectral density of the transmitted signals [1]

$$A_{rk}^2 = A_{rj}^2 = A_r^2 \quad \text{with any } k \text{ and } j$$

FOOTNOTE [1]. Recall that this character of spectral density in broadband signals is desirable for a decrease in the width of the range of powerful correlation (decrease in the time of correlation). ENDFOOTNOTE.

Then value R_i assumes the form

$$R_i(\Delta t_i) = \frac{P_{oi}}{\mu_i^2 FT} \sum_{k=k_1}^{k_2} \cos k\omega_0 \Delta t_i = \frac{P_{oi}}{\mu_i^2 FT} \frac{\sin \Delta\omega \frac{\Delta t_i}{2}}{\sin \omega_0 \frac{\Delta t_i}{2}} \cos \omega_{op} \Delta t_i, \quad (3.2.8)$$

where $P_{oi} = \frac{\mu_i^2}{2} \sum_{\kappa=\kappa_1}^{\kappa_2} A_{r\kappa}^2 = \frac{\mu_i^2 A_r^2}{2} FT$ - the power of the signal, which comes in on i-th ray/beam; $\Delta\omega$ = FT ω_0 - the bandwidth of signal; $\omega_{op} = \frac{k_2 + k_1}{2}\omega_0$ is the medium frequency of the spectrum of signal.

The exemplary/approximate form of the dependence R_i on Δt_i with FT >> 1 is shown in Fig. 3.2.2a. The character of a change in the envelope (3.2.8) from relative time of the disagreement/mismatch

of the transmitted and reference signals when $-\frac{T}{2}<\Delta t_i<\frac{T}{2}$ is shown in Fig. 3.2.3 [2].

FOOTNOTE [2]. During change Δt_i limits $-\infty<\Delta t_i<\infty$ the envelope (3.2.8) is function even and periodic with period of T. Figure 3.2.3 shows the curve/graph of this function only for one period. ENDFOOTNOTE.

As the parameter of curves are used the different values of the base of signal FT. As can be seen from figure, envelope has the fundamental maximum (range of powerful correlation), arrange/located about $\Delta t_i = 0$. The half of the width of this range is equal approximately to time of correlation τ_K, which composes value (see §2.5)

$$\tau_\kappa=\frac{1}{F}. \qquad (3.2.9)$$

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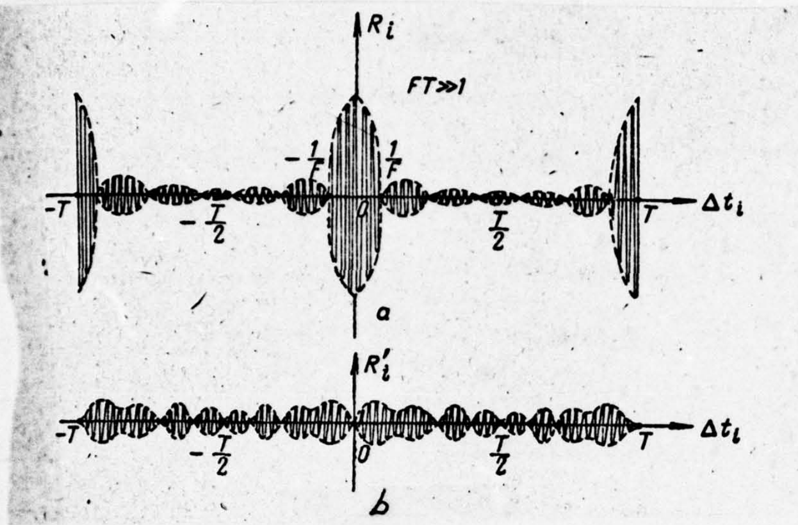
$$-\frac{T}{2}<\Delta t_i<-\frac{1}{F}$$

Besides the fundamental maximum the envelope has with FT $\geqslant$ 3 also a series of supplementary maximums, arrange/located in ranges and $\frac{1}{F}<\Delta t_i<\frac{T}{2}$. With an increase in the base of signal FT, i.e., with an increase in the band of signal frequencies F with the constant speed of transmission of information (T = const) decreases the width of the range of powerful correlation and descends the level of supplementary maximums. In this case as it was shown into §2.5, for FT >> 1 level of the latter it is considerably less than the fundamental maximum: greatest of the supplementary maximums does not exceed 20o/o of the level of the fundamental maximum, and remaining supplementary maximums have even smaller value. Thus, output potential of the correlator, matched with signal, depends substantially on the value of the utilized band of frequencies F.

Fig. 3.2.2. Page 132.



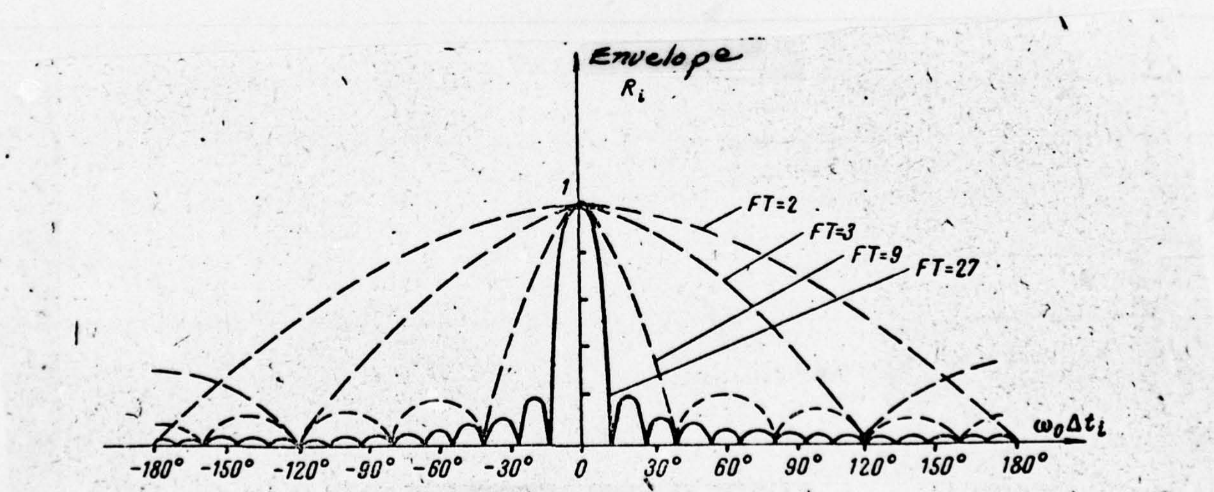


Fig. 3.2.3.

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In usual narrow-band communicating systems band F is small, the width of the range of powerful correlation is great. Therefore output potential of correlator will be created not only by the signal, which came on the synchronized ray/beam $(\Delta t_i = 0)$, but also by the signals, which came on other paths of propagation with time of disagreement/mismatch Δt_i, which are changed over wide limits. As a result of the interference of such signals appear selective fadings.

In the case of using broadband signals the width of the range of powerful correlation considerably decreases. If we in this case make a band sufficiently wide, then output potential of correlator will be created in essence only by signal, which came on the synchronized ray/beam. All other signals, which came in the ray/beams, which anticipate/lead that which was synchronized delaying relative to it more than on $\tau_k = \pm \frac{1}{F}$ will create on the output/yield of multiplier only very small stresses.

Let us examine now terms of the type $R_i'(\Delta t_i)$ from (3.2.7), that determine stress from the incoming ray/beams on the mismatched correlator. The exemplary/approximate form of the dependence of value R_i' from relative time lag Δt_i the transmitted and reference signals on the mismatched correlator for case FT >> 1 is represented in Fig. 3.2.2b. As can be seen from figure, this dependence has complex, to a certain extent "noise-like" character. At the torque/moment of reading R_i' it is equal to zero. Signals, that came on other ray/beams, create stresses with t = T, different from zero. Their effect with a sufficient degree of accuracy can be approximated by normal fluctuating interference with the uniform spectrum in the band of frequencies F.

Isolation/liberation of one of the incoming ray/beams in the system, which uses broadband signals, can be realized also with the aid of matched filters. Let the input of the matched filter of Fig. 3.2.1b enter the input signal x (t), determined by expression (3.2.2). The pulse response of the matched with signal $z_r(t)$ filter takes the form

$$G(t)=az_r(T-t), \qquad (3.2.10)$$

where a - certain constant coefficient. Then output potential of filter at certain moment of time t in accordance with results §2.8 is

equal

$$u_{\text{вых}}(t)=\int_{-\infty}^{\infty} x(t')\,G(t-t')\,dt'= =a\int_{-\infty}^{\infty} x(t')\,z_r(T-t+t')\,dt'. \qquad (3.2.11)$$

At the torque/moment of reading t = T this stress with an accuracy to constant factor coincides with the value of short-term crosscorrelation function between received signal x (t) and the expected signal $z_r(t)$.

By substituting in (3.2.11) value of x (t) from (3.2.2), we will obtain

$$u_{\text{вых}}(t)=a\sum_{i=1}^{n}\mu_i R_{0i}(t)+a\int_{-\infty}^{\infty}\xi(t')z_r\times \times(T-t+t')\,dt', \qquad (3.2.12)$$

where

$$R_{ol}(t)=\int_{-\infty}^{\infty} z_r(t'-\Delta t_l)\,z_r(T-t+t')\,dt'. \quad (3.2.13)$$

Second term in (3.2.12) is the result of the effect of fluctuating interference ξ (t) on matched filter. First term is the response of filter to the transmitted signals, which come in on different ray/beams. It is the result of the superposition (imposition) of responses $R_{ci}(t)$ from each of the incoming ray/beams individually.

Let us examine in more detail components of the type $R_{ci}(t)$. Since the incoming signal is different from zero only in interval $\Delta t_i \leqslant t \leqslant T+\Delta t_i$, the pulse characteristic of filter is different from zero with $t+\Delta t_i-T \leqslant t' \leqslant t+\Delta t_i$. Then from (3.2.13) we obtain

$$R_{0i}(t) = \int\limits_{t+\Delta t_i-T}^{t+\Delta t_i} z_r(t'-\Delta t_i) z_r(T-t+t')\,dt'.$$

Let us replace in this expression the variable of integration, after designating $y = T - t + t'$. Then

$$R_{0i}(t) = \int\limits_{\Delta t_i}^{T+\Delta t_i} z_r(y)\, z_r(y-\tau)\,dy, \tag{3.2.14}$$

where $\tau = T - t + \Delta t_i$.

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By comparing expressions (3.2.14) and (2.3.12), we come to the conclusion that the dependence R_{ci} on τ it is with an accuracy to constant factor the short-term autocorrelation function of signal $z_r(t)$. Therefore at the uniform spectral density of the transmitted signals this dependence with an accuracy to constant factor coincides a by the autocorrelation function of form (2.5.7). The curve/graph of its envelope for case FT >> 1 is shown in Fig. 3.2.4a. As can be seen from figure, envelope has distinct maximum as the width of order 2/F with τ = 0.

The dependence of response R_{ci} on t has analogous character. However, when τ passes value from 0 to T, then t varies from $T+\Delta t_i$. to Δt_i. Therefore the maximum of the envelope of stress from i-th ray/beam on the output/yield of filter occurs at torque/moment $T+\Delta t_i$ (see Fig. 3.2.4a).

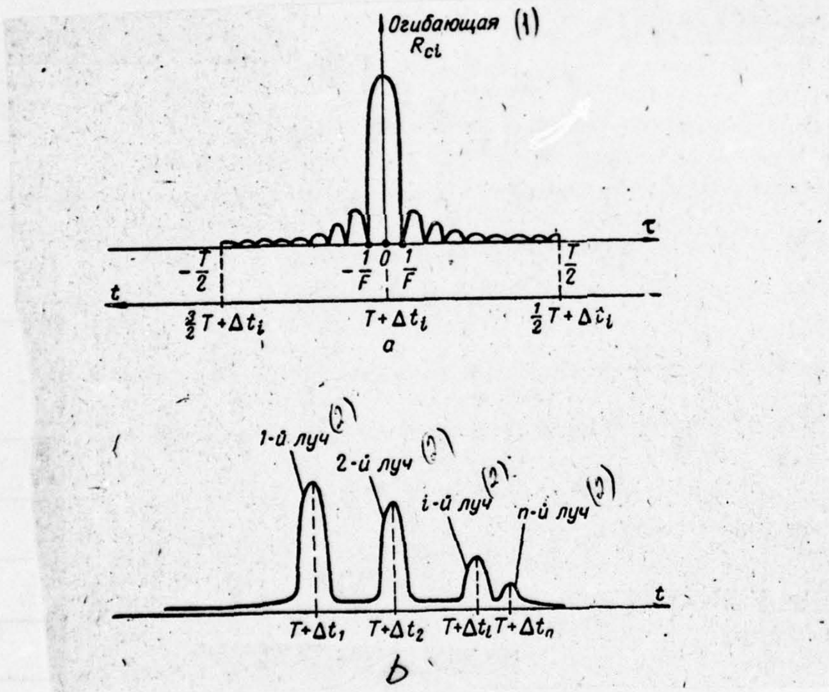
Огибающая (1)
Rci
τ
t
−T/2
−1/F
0
1/F
T/2
3/2 T+Δti
T+Δti
1/2 T+Δti
a
1-й луч (2)
2-й луч (2)
i-й луч (2)
n-й луч (2)
T+Δt1
T+Δt2
T+Δti
T+Δtn
b

Fig. 3.2.4.

Fig. 3.2.4.

Key: (1). Envelope. (2). 1-1 ray/beams. Page 136.

Thus, each of the incoming **ray/beams creates at** the output/yield of filter at the moments of time $T+\Delta t_i$, $i=1, 2, \ldots, n$ their maximum of **stress.** If path differences between adjacent ray/beams exceed value $\pm\frac{1}{F}$, then these maximums are not superimposed one on top of the other (Fig. 3.2.4b) and do not interfere, as it takes place for narrow-band signals. By the selection of sufficiently broad band F it is possible to effectively divide the incoming ray/beams. After combining the torque/moment of reading with one of them (by the first or greatest), it is possible to carry out a reception/procedure only of one ray/beam.

By discussing analogously, it is possible to show that the effect of the incoming ray/beams during the transmission of signal $z_r(t)$ to the filter, matched with another signal $z_l(t)$ takes the form

$$u'_{вых}(t)=a\sum_{i=1}^{n}\mu_i R'_{ci}(t), \tag{3.2.15}$$

where

$$R'_{ci}(t)=\int_{\Delta t_i}^{T+\Delta t_i} z_r(y-\tau)\,z_i(y)\,dy.$$

Each term in this expression is with an accuracy to constant factor the mutually correlated function between signals $z_r(t)$ and $z_i(t)$. When using broadband signals the dependence R'_{ci} on τ has fairly complicated "noise" character. The graph/diagram of this dependence for the signals orthogonal in the intensive sense is similar curve/graph, shown in Fig. 3.2.2b, if we along the axis of abscissas instead of Δt_i plot τ. At the moment of time $t=T+\Delta t_i\ (\tau=0)$ value R_{ci}

is equal to zero and is different from zero for the ray/beams, which come in at other moments of time, i.e., output potential of filter in this case it is created by all ray/beams, except that which is taken. For practical calculations it with a sufficient degree of accuracy can be approximated by normal fluctuating interference.

Thus, in the communicating system, which uses broadband signals, appears the possibility of the effective separation of the incoming ray/beams with subsequent processing one of them.

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Therefore the decisive diagrams of the receptors of such systems can be realized in essence just as the diagrams of single-ray reception/procedure, examined into §2.7 and 2.8. In this case is feasible both coherent and incoherent reception/procedure with the realization of the decisive diagrams either on multipliers (see Fig. 2.7.2 and 2.7.4), or on matched filters (Figs. 2.8.5 and 2.8.7). The output voltages in the branches of processing the indicated diagrams are proportional at the torque/moment of reading T to values X_r, r = 1; 2, with coherent reception/procedure and V_r with incoherent. The

values X_r and V_r in the branches, matched with the transmitted signal $z_r(t)$ are determined by the taken in the synchronized ray/beam signal $\mu_i z_r(t)$ and fluctuating interference ε (t), but in the mismatched branches - only by interference. The latter consists of the normal fluctuating noise of the channel of communication/connection and supplementary noise, formed because of the action of other ray/beams.

The important condition of the isolation/liberation of useful signal is its synchronization with the signal of reference oscillator.

Let us examine requirements for the synchronization of the adopted ray/beam in such diagrams. Figure 3.2.5a depicts to the dependence of the stress at torque/moment t = T at the output/yield of the matched with the transmitted signal branch of processing in the decisive diagram on multipliers from Δt - relative superimposition at time of the adopted and reference signals. Figure 3.2.5b shows the dependence of the output voltage in this branch from the current time t. To relative superimpositions 0, Δt_1, Δt_2 correspond the values of the output voltages, proportional X_r, $X_r^{(1)}$, $X_r^{(2)}$ in the case coherent and V_r, $V_r^{(1)}$, $V_r^{(2)}$ in the case of noncoherent

reception/procedure.

As can be seen from figures, the absence of the synchronism between the torque/moment of the arrival of the transmitted signal and the beginning supporting/reference leads to changes in the output stresses at the torque/moment of reading. However, in the case of coherent reception/procedure these changes are more considerable: if one and the same superimposition of Δt_2 in incoherent reception/procedure corresponds value $V_r^{(2)}$, which differs little from V_r, then with coherent reception/procedure it occurs

X_r

even a sign change of value $X_r^{(2)}$. In the general case the dependence on Δt is oscillating with period $T_{cp}=\frac{2\pi}{\omega_{cp}}$, where ω_{cp} is the medium frequency of the spectrum of the workable signal. Page 138.

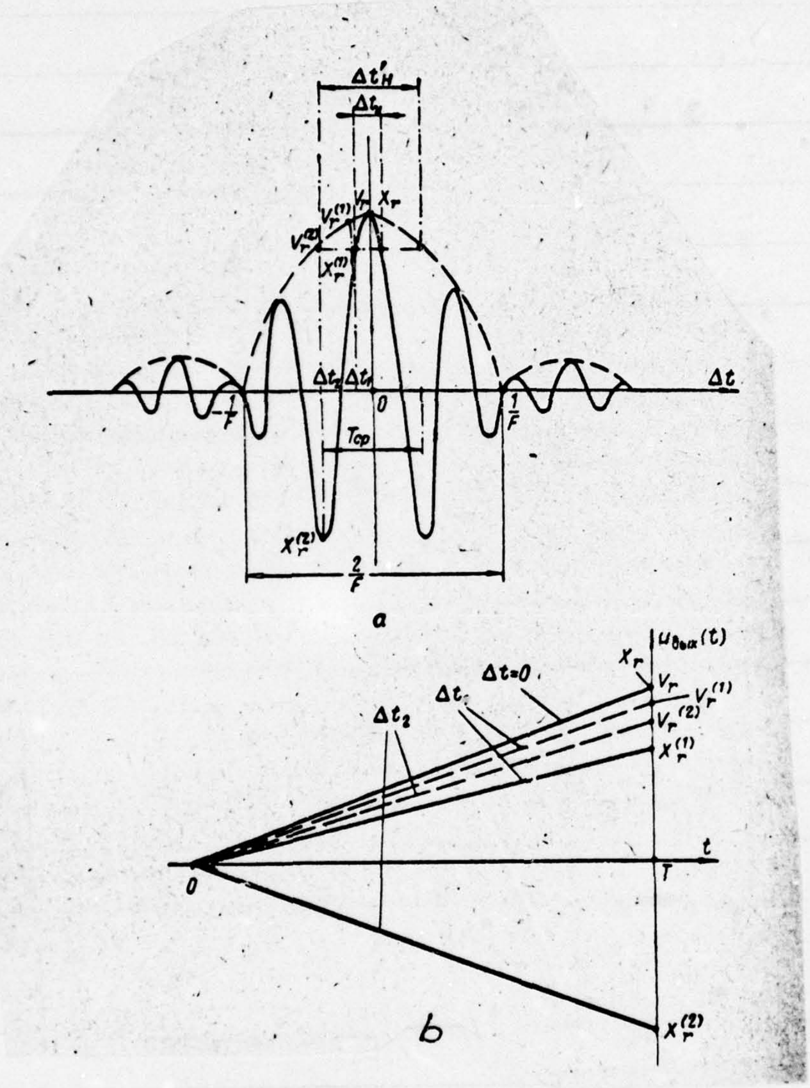

Fig. 3.2.5.

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Essential change V_r occurs when Δt commensurable with value 1/F. In order that the indicated changes in values X_r and V_r would be sufficiently small and did not cause the decreases in the freedom from interference, permissible the superimposition of the workable and reference signals must satisfy the follcwing conditions:

$$\left.\begin{aligned} &\frac{\Delta t_{\text{н}}}{2} \ll \frac{T_{\text{ср}}}{4} \text{ — with coherent reception/procedure;} \\ &\frac{\Delta t'_{\text{н}}}{2} \ll \frac{1}{F} \text{ — with inccherent reception/procedure.} \end{aligned}\right\} \quad (3.2.16)$$

Consequently, in the case of coherent reception/procedure synchronization must be realized with an accuracy to the small fraction of a period of the medium frequency of the spectrum of signal, which in a number of cases can turn out to be hard to produce. With the incoherent reception/procedure of the condition of synchronization considerably they are facilitated, since $T_{\text{ср}} \ll \frac{1}{F}$. The accuracy of reading of the output voltages at torque/moment t = T, as can be seen from Fig. 3.2.5b in diagrams on multipliers it must be the order of the portion of value T.

Figure 3.2.6 shows the dependences of output voltage from the current time t in the diagrams of coherent and incoherent reception/procedure with matched filters. From the figure one can see that in such diagrams of requirement for the permissible displacement of fiducial mark at torque/moment $T+\Delta t_i$ are analogous to conditions (3.2.16).

Besides those which were indicated, for isolation/liberation of one of the incoming ray/beams in broadband systems can be used other versions of the decisive diagrams. Specifically, is possible the construction of the diagrams both of coherent and incoherent reception/procedure, which is based on the principle of the so-called synchronous heterodyning.

Figure 3.2.7 depicts the functional diagram of synchronous heterodyning for the case of the incoherent reception of binary signals. The diagram contains two reference oscillators (heterodyne), the forming signals z_{on1} and z_{on2}, which must be strictly synchronized with the workable ray/beam and differ from the transmitted signals in terms of shift/shear by the value of intermediate frequency $\omega_{пр1} = k_{пр1}\,\omega_0$

and $\omega_{np2} = k_{np2}\,\omega_0$ respectively.

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Received signal x (t) enters two multipliers (mixer), to which simultaneously are introduced the reference signals z_{on_r}. The output voltages of multipliers enter then the kinematic filters - ducts of the high quality, which have the passbands of order 2π/T and resonance frequencies ω_{np1} and ω_{np2} respectively. For processing each cell/element of signal the filters must be equipped by the devices, which make it possible periodically after reading at torque/moment t = T to instantly extinguish the natural oscillations in them. It is possible to show that the output potentials of filters up to the torque/moment of reading contain the only components of intermediate frequencies ω_{np1} and ω_{np2}. The instantaneous values of these voltage/stresses at the torque/moment of reading with an accuracy to constant factor coincide with values X_r, and amplitude (envelopes) them - with values V_r^*. In the case of the incoherent reception of the amplitude of the output voltage/stresses of filters they are isolated by envelope detectors and they enter comparison circuit, where conducts the selection of that signal $z_r(t)$, for which value V_r is more.

From the viewpoint of the possibility of isolation/liberation of one of the incoming ray/beams, the freedom from interference and the conditions of synchronization, the examined diagram is completely equivalent to the diagrams of incoherent reception on multipliers and matched filters. At the same time from the comparison of Figs. 2.7.4 and 3.2.7 it is evident that the diagram of synchronous heterodyning contains the smaller number of functional subassemblies. As concerns diagram with matched filters, the realization of the filter, matched with the signal of complex form, by itself represents fairly complicated problem. Therefore the use of a synchronous heterodyning can ensure in a number of cases of advantage during the technical realization of receptors, especially during processing and the useful use of a series of the incoming ray/beams. The principles of synchronous heterodyning found use in the broadband communicating system of the type "Rake", examined in §3.5.

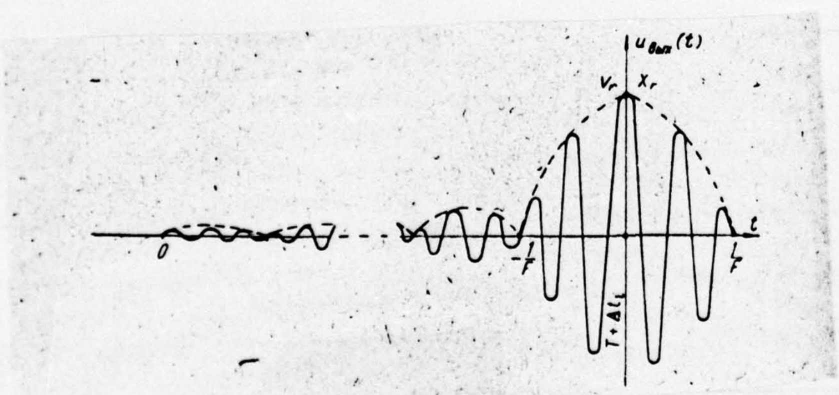
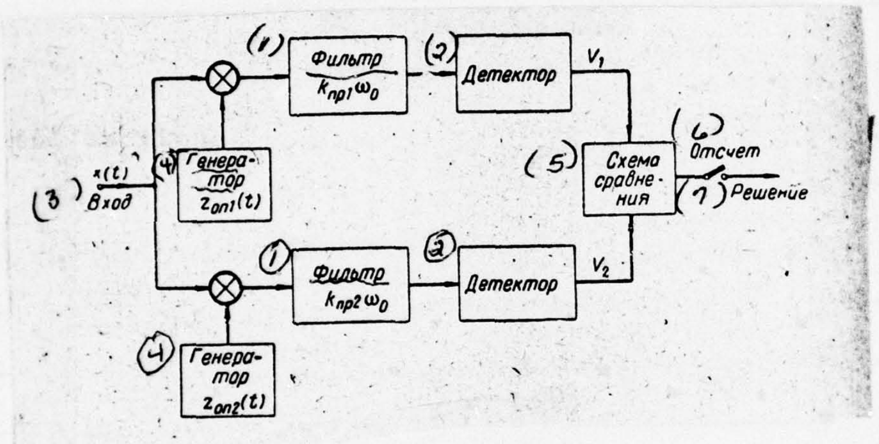


Fig. 3.2.6.

Fig. 3.2.7.

Key: (1). Filter. (2). Detector. (3). Input. (4). Generator. (5). Comparison circuit. (6). Reading. (7). Solution.



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§3.3. Principle of the use of several ray/beams and the decisive diagrams for a multiple-pronged reception.

The diagrams of isolation/liberation of one of the incoming ray/beams in broadband communicating systems make it possible in principle to get rid of the effect of echo and selective fadings on the reception of signals in channels with multiple-pronged

propagation. However, the correctness of the transmission of information in such systems is determined only by energy of one workable ray/beam.

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At the same time each of the incoming ray/beams includes useful information. In connection with this does arise the question: it is not possible whether, utilizing knowledge of the structure of multiple-pronged signal, so to process it in receptor in order to obtain total energy of entire signal and to raise thereby the correctness of the reception of information? Actually arises the question concerning the construction of the decisive receiver circuit, in the best way matched with multiple-pronged signal.

Answer/response to this question is obtained during the last/latter decade in the investigations, made in a series of Soviet and foreign works, for example [36, 24] and others. From these works it follows that optimum in the sense of ideal observer's criterion processing multiple-pronged signal in receptor can be based on the use of broadband signals and assumes, in the first place, the

isolation/liberation of each of the incoming ray/beams and, in the second place, their subsequent addition in accordance with the available with data acquisition on the structure of the multiple-pronged communication channel.

For the explanation of the principle of the use several the incoming into the point of reception ray/beams let us turn to diagram in Fig. 3.3.1. In this figure is shown the decisive diagram of the receptor of binary communicating system, which processes the multiple-pronged signal of the form

$$x(t)=\sum_{i=1}^{n}\mu_i z_r(t-\Delta t_i)+\xi(t), r=1;\ 2, \qquad (3.3.1)$$

where $z_r(t)$ are the utilized for a transmission information broadband signals; n - the number of incoming ray/beams; μ_i and Δt_i - transmission factor and the time lag of I-th ray/beam; $\xi(t)$ - the additive fluctuating interference. As can be seen from figure, the circuits of processing signals z_1 (t) and z_2 (t) consist of the identical parallel branches whose number is equal to the amount of workable ray/beams. Each branch is correlator and switches on

supporting voltage generator $\mu_i z_r(t-\Delta t_i)$, multiplier and integrator. Reference oscillators are synchronized with the appropriate workable ray/beams, i.e., the beginning of the cell/element of the signal of the generator, synchronized with the first incoming ray/beam, it coincides with the delay time in the first ray/beam Δt_1, the second in the delay time in the second ray/beam Δt_2 and in the, etc. Page 143.

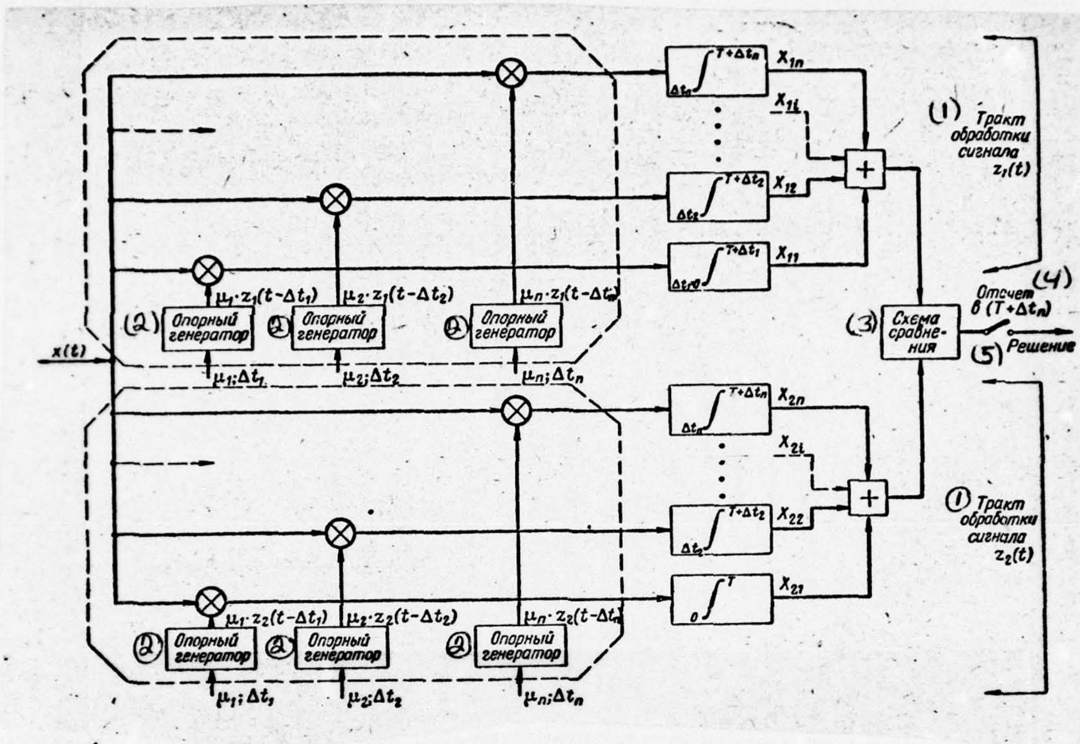
x(t)
(2) Опорный генератор
$\mu_1 \cdot z_1(t-\Delta t_1)$
$\mu_2 \cdot z_1(t-\Delta t_2)$
$\mu_n \cdot z_1(t-\Delta t_n)$
$\mu_1; \Delta t_1$
$\mu_2; \Delta t_2$
$\mu_n; \Delta t_n$
$\mu_1 \cdot z_2(t-\Delta t_1)$
$\mu_2 \cdot z_2(t-\Delta t_2)$
$\mu_n \cdot z_2(t-\Delta t_n)$
X_{1n}
X_{1i}
X_{12}
X_{11}
X_{2n}
X_{2i}
X_{22}
X_{21}
(1) Тракт обработки сигнала $z_1(t)$
(1) Тракт обработки сигнала $z_2(t)$
(3) Схема сравнения
(4) Отсчет в $(T+\Delta t_n)$
(5) Решение

Fig. 3.3.1.

[redacted]

Key: (1). Circuit of processing signal. (2). Reference oscillator. (3). Comparison circuits. (4). It will swell. (5). Solution. Page 144.

Subsequently for simplicity let us count off the time lag of ray/beams in formula (3.3.1) from the torque/moment of input process of the receptor of the first ray/beam, so that $\Delta t_1 = 0$, and the others Δt_i are different from zero and are positive, whereupon is satisfied condition $0<\Delta t_2<\Delta t_3<\ldots<\Delta t_n$.

The voltages of reference oscillators and the adopted ray/beams enter the multipliers, and then the integrators. The work of integrator in each branch of processing is synchronized on time with the work of the corresponding reference oscillator. Time of integration is equal to the duration of the cell/element of signal T. At torque/moment $T+\Delta t_i$ the termination of the cell/element of the signal of I-th reference oscillator conducts the reading of voltage X_{ri} on the output/yield of I-th integrator and its supply into adder, whereupon the diagram of integration by means of the rapid jettisoning of voltage is prepared for processing the following

cell/element of signal.

The output voltages of the integrators of the circuits of processing signals z_1 (t) and z_2 (t) store/add up in adders and enter the comparison circuit. In this diagram at the moment of time $T+\Delta t_n$ the termination of processing the latter from the come ray/beams are compared the output voltages of the circuits of processing and on larger of them it is accepted solution to the reception of one signal or the other.

The dialing/sets of correlators in the circuits of processing signal are in essence the device, which divides those which come in to luy. The isolation/liberation of each ray/beam in this device when using broadband signals is based on what the correlation function of such signals has very narrow range of powerful correlation. In §3.2 it was shown, that the correlator, synchronized with any ray/beam, reacts to the incoming signal with the band of frequencies F only in time interval ± 1/F relative to the torque/moment of synchronization. In this case the voltage on its output/yield at the torque/moment of the termination of the cell/element of signal is proportional to value X_r - short-term crosscorrelation function between that which is taken in ray/beam and reference signals (see Fig. 3.2.5).

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Consequently, if we select the frequency band by the so wide in order that value 1/F would be less than the time lag between ray/beams and to utilize a series of the correlators, each of which was synchronized with one of the incoming ray/beams, then it is possible to isolate all the incoming ray/beams. Of this is the first development stage of multiple-pronged signal - the stage of the isolation/liberation of each of the incoming ray/beams, of realized in diagram in Fig. 3.3.1 dialing/set of correlators.

Figure 3.3.2 shows output potentials of the integrators of the correlators any from the circuits of processing signal. At the moments of time T, T + Δt_2, ..., $T+\Delta t_n$ the output potentials of the correlators of the branches of processing are proportional to short-term crosscorrelation functions X_{rt} only between the signal of the corresponding ray/beam and synchronized with it reference oscillator.

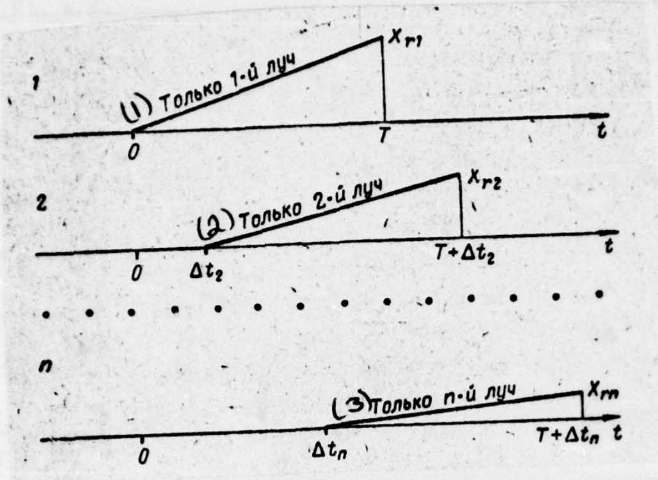
1
(1) Только 1-й луч
x_{r1}
0
T
t
2
(2) Только 2-й луч
x_{r2}
0
Δt_2
$T+\Delta t_2$
t
n
(3) Только n-й луч
x_{rn}
0
Δt_n
$T+\Delta t_n$
t

Fig. 3.3.2.

~~Page 3.3.2:~~

Key: (1). Only 1-1 ray/beams. (2). Only 2nd beam. (3). Only 7th ray/beam. Page 146.

In the diagram in question will be divided all the incoming ray/beams, if time of their time lag satisfies the condition

$$\Delta t_i > \frac{1}{F}. \qquad (3.3.2)$$

As a result of this separation is removed the pileup effect of the adjacent cell/elements of signal on each other, i.e., is removed the phenomenon of echo. However, one separation of ray/beams still insufficiently for obtaining in the adders of the energy of entire multiple-pronged signal. For the isolation/liberation of total energy of the incoming signal is necessary the second stage of its processing - the phasing and the amplitude weighing of the incoming

ray/beams in accordance with the available data on the structure of the multiple-pronged communication channel. Such data are the values of the transmission factors and delay time in each ray/beam. The addition of values μ_i and Δt_i for an I-th ray/beam is realized in Fig. by 3.3.1 into the schematics of reference oscillators and is conditionally shown by rifleman/pointers. The devices, which determine the values μ_i and Δt_i, in figure are not shown.

Let us explain the sense of the operations of phasing and amplitude weighing. As it was shown in chapter 1, the signals, coming in on separate ray/beams, are characterized by the different initial phases of high-frequency filling, which depend on the time lag of I-th ray/beam Δt_i, and by the different amplitudes, determined by transmission factors μ_i. Specifically, as a result of the direct "disordered" addition of such signals appear selective fadings.

It is obvious that for using an energy of entire multiple-pronged signal it is necessary first of all, in order that the signals, which come in on separate ray/beams, would store/add up with identical phases. Satisfaction of this condition composes the sense of the operation of phasing. As it was shown in §3.2, the absence of the synchronism between the torque/moment of the arrival

of any ray/beam and the beginning of the cell/element of the signal of the corresponding reference oscillator can lead to substantial changes in the output voltage of correlator up to a change in its sign (see Fig. 3.2.5). In order that changes in the output voltage of correlator would be insignificant, necessary the synchronization of reference oscillator with an accuracy to the small fraction of a period of the medium frequency of the spectrum of signal. In this case they speak about synchronization to accuracy "up to the phase of high-frequency filling of signal". In diagram in Fig. 3.3.1 phasing of the workable ray/beams will be made when each reference oscillator will be synchronized with the incoming ray/beam up to the phase of high-frequency filling of signal.

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In this case the output voltages of correlators in the circuits of processing will be sinphase and, entering the adders, they store/add up arithmetical, but it is not vector (Fig. 3.3.3). In other words, as a result of the operation of phasing is realized the coherent addition of the signals, which come in in separate ray/beams. It is natural that as a result of phasing completely is removed the selective character of fadings.

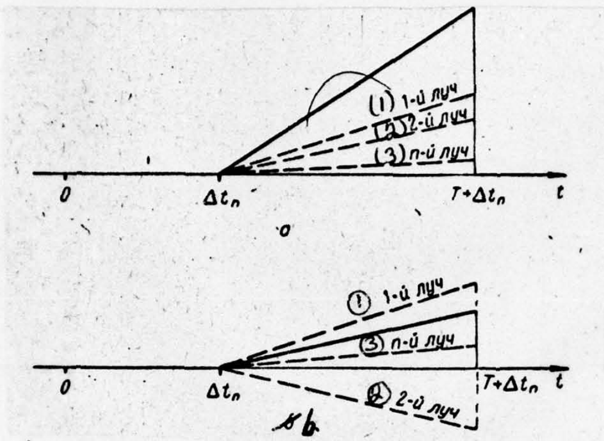

Fig. 3.3.3.

Key: (1). y-and ray/beam. (2). 2-and ray/beam. (3). pi are ray/beam.

Besides phasing optimum processing multiple-pronged signal assumes also conducting the operation of the amplitude weighing of each of the incoming ray/beams. The target/purpose of amplitude weighing lies in the fact that, in order to increase the energy of entire workable signal with respect to the spectral density of fluctuating interference and to decrease thereby the probability of the appearance of errors in communicating system.

Of the lots of the workable ray/beams in the isolation/liberation of total energy of multiple-pronged signal it is different. The role of ray/beams with large transmission factors is great, at the same time the ray/beams with small μ_i are located in essence under the effect of interferences. [redacted]

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The operation of amplitude weighing provides the best isolation/liberation most intense of the adopted ray/beams and

attenuates ray/beams with low transmission factors, whereupon the weakening must be the more greater, the the less corresponding value μ_i. In the same way as this occurred in matched filter, as a result of amplitude weighing as "are emphasized" the most intense "components" of multiple-pronged signal. Is realized the operation of weighing by means of the multiplication of the output voltages of the correlators, which enter the input of adders, on the appropriate weight coefficient (factor) μ_i. Moreover it is indifferent, at which point of correlator to realize is functional this operation: prior to the input of multiplier from received signal or from reference oscillator, after multiplier prior to the input of integrator, after integrator. In diagram in Fig. 3.3.1 amplitude weighing is made prior to the input of multipliers from reference oscillators by means of the multiplication of reference signals $z_r(t-\Delta t_i)$ by weight factor μ_i.

The examined diagram explains the principle of optimum processing multiple-pronged signal; however, its practical realization for the isolation/liberation of the sufficiently large number of ray/beams is difficult. Actually, the amount of incoming into the point of reception ray/beams and time lag between them they depend on the state of the channel of propagation and randomly change in time. For the confident isolation/liberation of each of these ray/beams it is necessary with high accuracy continuous in time to

adjust slightly the synchronization of reference oscillators to strict preservation/retention/maintaining of identical phase relationship/ratios both between the reference oscillators in each circuit of processing and between the generators of the circuits of signals z_1 (t) and z_2 (t). Therefore the presence of the sufficiently large number of reference oscillators and integrators with complex individual timing mechanisms makes diagram in Fig. 3.3.1 extremely bulky. The practical realization of this diagram can turn out to be advisable when either the number coming in or the number of workable most powerful ray/beams relatively small and does not exceed two-three.

Optimum processing the totality of all broadband signals, which enter the input of receiver under conditions of multiple-pronged propagation, can be functionally substantially simplified, if the synchronization of each correlator is carried out, after introducing time lag not into the signals of reference oscillators, but into the adopted ray/beams.

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The sense of the introduction of this delay can be explained as follows. Let the input of receptor enter n of ray/beams with different at times time lag Δt_i, i = 1, 2, ..., n. Let us accept for the zero time reference the torque/moment of arrival at the receiver of the first ray/beam ($\Delta t_1 = 0$). In the case of diagram with the synchronization of reference oscillators under each of the incoming ray/beams the output voltages of correlators in the branches of any circuit of processing signal (z_1 (t) or z_2 (t)) occupy in time consecutive positions at torque/moments T, T + Δt_2, ..., $T+\Delta t_n$, as this is shown in Fig. 3.3.2, and, also consecutively entering the adder, they store/add up in it.

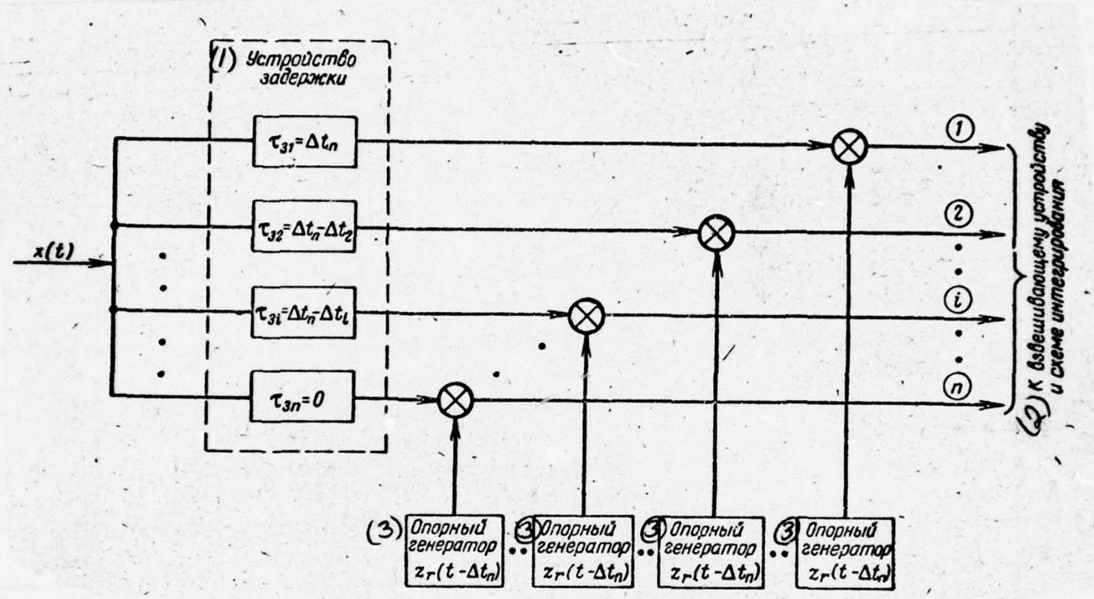
Let us introduce delay for a received signal with the aid of the delay units, which stand at the input of correlators in the branches of the circuits of processing signals z_1 (t) and z_2 (t) (Fig. 3.3.4). Moreover input multiple-beam signal in the branch of processing the first ray/beam let us delay for a period $\tau_{31}=\Delta t_n$, in the branch of processing the second ray/beam - for a period $\tau_{32}=\Delta t_n-\Delta t_2$, in I-th branch - for a period $\tau_{3i}=\Delta t_n-\Delta t_i$ and finally into the branch of processing the last/latter ray/beam let us feed signal not delayed. simultaneously we synchronize all reference oscillators at the torque/moment of the arrival of the n ray/beam. Then in the first branch of processing the synchronized and reference oscillator

render/shows the first of the incoming ray/beams, in the second - the second ray/beam, etc (Fig. 3.3.5). These ray/beams will be isolated at the output/yield of correlators at one and the same moment of ~~time~~ $T+\Delta t_n$. Figure 3.3.5 they shows by solid lines. The unsynchronized ray/beams in each branch of processing (Fig. 3.3.5 they shows conditionally by dotted line) create on the output/yield of correlators only the insignificant voltages, which can be set/assumed by the virtually equal to zero, if path differences of ray/beams satisfy condition (3.3.2).

In the case in question the beginning of the cell/elements of the signals of reference oscillators as in the circuit of processing signal z_1 (t), so also in circuit z_2 (t) must correspond to the torque/moment of the arrival of the last/latter ray/beam Δt_n. Then there is no need for for the series of reference oscillators with individual timing mechanisms, and sufficient to provide for each circuit of processing on one reference oscillator with synchronization at torque/moment Δt_n, **Page** 150.

Fig. 3.3.4.

Key: (1). Delay unit. (2). To the weighhing device and the diagram of integration. (3). Reference oscillator. Page 151.



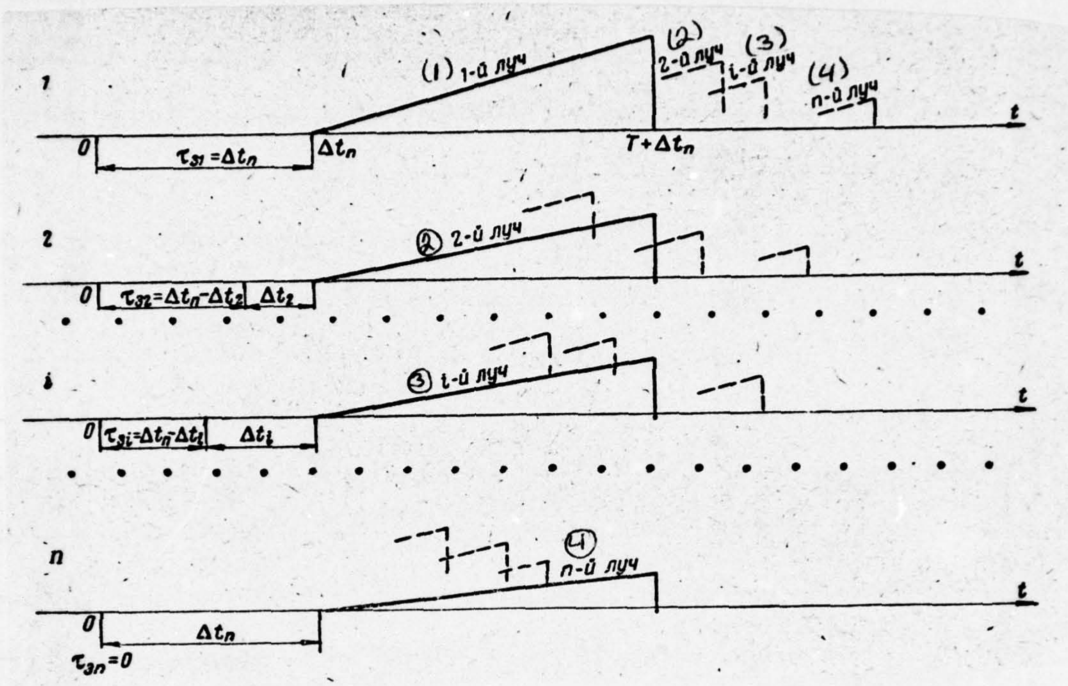

1
(1) 1-й луч
(2) 2-й луч
(3) i-й луч
(4) n-й луч
0
$\tau_{з1}=\Delta t_n$
Δt_n
$T+\Delta t_n$
t
2
(2) 2-й луч
$\tau_{з2}=\Delta t_n-\Delta t_2$
Δt_2
i
(3) i-й луч
$\tau_{зi}=\Delta t_n-\Delta t_i$
Δt_i
n
(4) n-й луч
Δt_n
$\tau_{зn}=0$

Fig. 3.3.5.

[redacted].

Key: (1). 1-1 ray/beams. (2). 2-1 ray/beams. (3). the i ray/beam. (4). pi are ray/beam. Page 152.

Of course, in this case the operation of the weighing of the adopted ray/beams already it is not possible to realize by multiplication of the signals of reference oscillators of weight coefficients μ_i. However, with identical success the weighhing device can be supplied at the output/yield of the multipliers of the correlators of each circuit of processing. Furthermore as it is not difficult to see, time of integration of the workable ray/beams with the method of the synchronization of received signals in all branches of processing is included within limits from Δt_n to $T+\Delta t_n$. Therefore, instead of summarizing signals after separate integrators, it is possible to use one common/general/total integrator in each of the circuits of processing signals z_1 (t) and z_2 (t). For this it is necessary to change the order of performance of the operations of addition and integration, which is unprincipled, since they are linear.

Finally, as the delay unit in the received signals it is expedient to utilize one delay line, which has common/general/total

input and separate output/yields for the correlators of the circuits of processing signals z_1 (t) and z_2 (t). The removal/outlets of line must be established/installed in order to provide the values of the indicated above delays τ_{3i}. However, as already mentioned, number of incoming to receptor ray/beams and time of their time lag change randomly, which requires in turn, of the random values τ_{3i}. Therefore confident processing all indicated ray/beams can be conducted, after entering as follows. Since single correlator reacts to the incoming signal only in time interval $\pm$ 1/F relative to the torque/moment of synchronization, the range of the possible time lags incoming ray/beams must be overlapped by the series of the correlators, each of which is synchronized with a consecutive increment in the time lag for value 1/F. Accordingly the delay line must have a series of output removal/outlets, arrange/located consecutively through time intervals 1/F. The last/latter removal/outlet must provide the delay, not less than time lag Δt_n the last/latter from those which come in for input receiver of ray/beams.

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Thus, when using a delay in the received signal the diagram of optimum processing the multiple-pronged signal of Fig. 3.3.1

equivalently converted into the diagram, presented in Fig. 3.3.6. As can be seen from figure, diagram consists of the line of the delay, series of correlators in the circuits of processing signals z_1 (t) and z_2 (t), the devices of weighing, adders (common/general/total "busbar" at the output/yield of the devices of weighing) and comparison circuits. The delay line provides a delay in received signals x (t) of form (3.3.1). Its removal/outlets are arranged consecutively through time intervals 1/F. The latter from removal/outlets corresponds to a signal delay x (t) for a period, not smaller than the time lag of the last/latter ray/beam Δt_n, characteristic for this channel of propagation. Correlators in the circuits of processing signals switch on the series of multipliers, the common for each circuit reference oscillator, synchronized with torque/moment Δt_n up to the phase of high-frequency filling of received signal, and the common for each circuit integrator, which enters the signal from the output/yield of adder. The device of weighing realizes multiplication of each of the taken ray/beams by the weight factor, equal to the transmission factor of this ray/beam. Thereby during the release of energy of entire multiple-pronged signal are emphasized the most intense ray/beams and are attenuate/weakened ray/beams with small μ_i. At the output/yield of those removal/outlets of the device, where the adopted ray/beams are absent, voltage is equal to zero.

Let us note that during practical realization as this will be shown below, to the device of weighing must be entrusted also the functions of the supplementary phase correction of the adopted ray/beams. The need for this is caused by the dephasing of ray/beams due to an error in the installation of the removal/outlets of delay line and because the time lag between ray/beams in real channel is accurate not multiply to time interval 1/F, although the latter can be selected by sufficiently small.

The output voltages of the integrators of the circuits of processing enter the comparison circuit, in which at the moment of time $(T+\Delta t_n)$ on larger of them is accepted the solution to the reception of signal z_1 (t) or z_2 (t). Page 154.

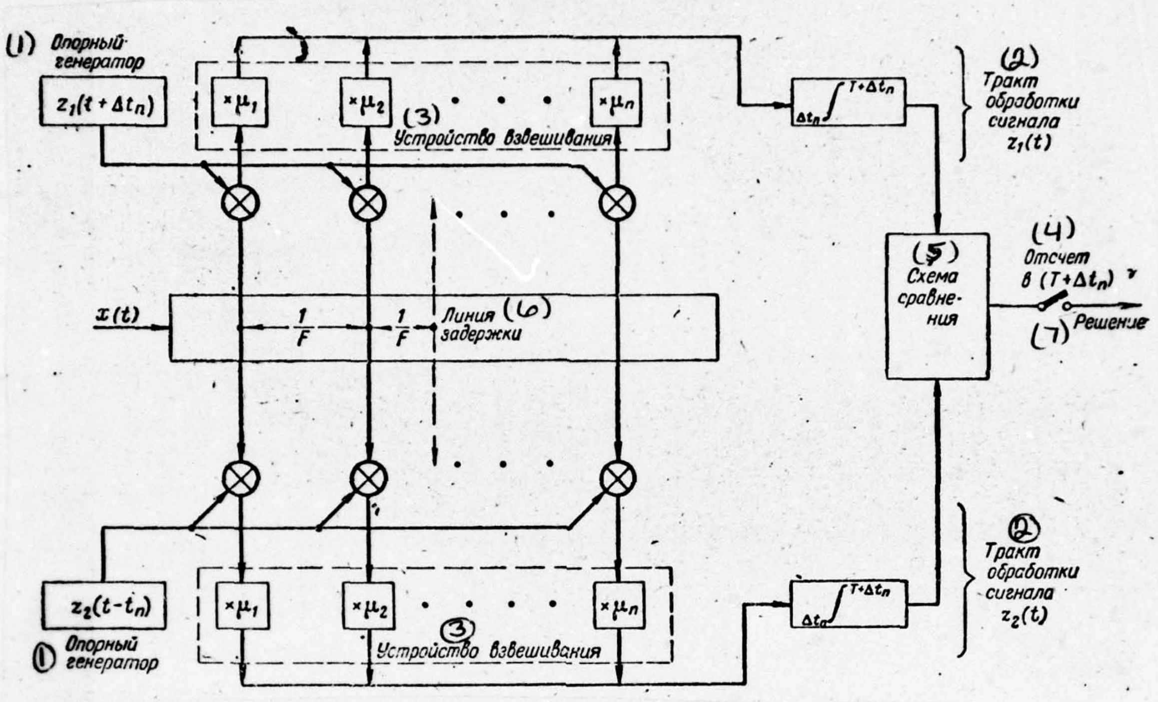
(1) Опорный генератор
$z_1(t+\Delta t_n)$
$\times\mu_1$
$\times\mu_2$
$\times\mu_n$
(3) Устройство взвешивания
Δt_n $\int$ $T+\Delta t_n$
(2) Тракт обработки сигнала $z_1(t)$
(5) Схема сравнения
(4) Отсчет в $(T+\Delta t_n)$
(7) Решение
$x(t)$
$\frac{1}{F}$
$\frac{1}{F}$
Линия (6) задержки
$z_2(t-t_n)$
(1) Опорный генератор
(3) Устройство взвешивания
(2) Тракт обработки сигнала $z_2(t)$

Fig. 3.3.6.

[redacted]

Key: (1). Reference oscillator. (2). Circuit of processing signal. (3). Device of broadcasting. (4). Reading. (5). Comparison circuit. (6). Pine of delay. (7). Solution. Page 155.

After tracing sequence of operations, executed by the decisive diagrams in Fig. 3.3.1 and of Fig. 3.3.6, it is possible to say that both these diagrams realize coherent reception with the coherent addition of the broadband signals, which came in different ray/beams. Of course, from the viewpoint of freedom from interference they are equivalent. Are identical also requirements in these diagrams with respect to the high (up to the phase of high-frequency filling of signal) accuracy of the synchronization of reference oscillators. However, diagram with a delay in the received signal is simpler than diagram with a signal delay of reference oscillators. In spite of the increased number of multipliers and the connected with this complication of the device of weighing in diagram with a delay in the received signal there is no series of integrators and, which is especially important, the series of reference oscillators with individual timing mechanisms.

The examined decisive diagrams explain the principle of the construction of the optimum in the sense of Kotelnikov's criterion (minimizing the composite probability of the error of piece-by-piece reception) receptors for the communication channels with normal fluctuating interference and sufficiently slow changes in the state, when the values of the transmission factors and initial phases of signals in ray/beams can be measured and known with reception.

Is at present a series of Soviet and foreign works, dedicated to the analysis of the optimum decisive diagrams for channels with multi-beam characteristics with the different volume of information relative to the parameters of these channels and the different character of the affecting interferences. Specifically, in work [36] it is indicated about the possibility of the construction of receptor in the form of diagram in Fig. 3.3.1, is given the rule of the solution of this diagram and modification of device. In work [24, chapter 9] on the base of the equivalent model of multiple-pronged channel, in which are utilized the selective values of its pulse transient function, obtained the rule of solution for a diagram in Fig. 3.3.6. Due to unwieldiness these rules of solution here are not given. If necessary the reader can use the directly indicated above works.

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Of course, the diagrams of coherent reception with the coherent addition of the broadband signals, which came in different ray/beams, can be constructed just as in the case of single-ray reception, either on the base of matched filters or by applying a principle of synchronous heterodyning. In this last/latter case when using a delay in the received signal most simply it is possible to combine the operations of weighing and phasing of ray/beams with the measurement of their transmission factors.

For explaining the aforesaid let us turn to Fig. 3.3.7, in which are represented the version of the examined above diagram with a delay in the received signal, the using principle of synchronous heterodyning. The diagram in Fig. 3.3.7 contains delay line in the series of the removal/outlets through intervals 1/F for the circuits of processing signals z_1 (t) and z_2 (t). The total delay time in the line τ_3 is not less than the time lag of the latter from the incoming ray/beams Δt_n. Its input enters the adopted multiple-pronged signal x (t) of form (3.3.1) with medium frequency f. The removal/outlets of line are connected to the multipliers (mixers) of the series A, to which is supplied also the voltage from reference oscillators z_1 (t) or z_2 (t) for each circuit of processing signal respectively. The generators of reference signals are synchronized so that the beginning of their cell/element would coincide with the torque/moment, when on the last/latter output of line render/shows the beginning of the cell/element of the signal, accepted on the first of the incoming ray/beams.

For the realization of the principle of synchronous heterodyning, in the first place, the reference oscillators reproduce

the transmitted signals z_1 (t) and z_2 (t) with medium frequency $f-\Delta_1$, i.e., with the frequency shift of all their harmonic components with respect to medium frequency transmitted signals to value Δ_1. In the second place, the role of the integrators of correlators perform kinematic filters (see §3.2). Then from the last/latter pair of the multipliers of series A arise the voltages, which contain the component of frequency Δ_1, and with instantaneous values at the torque/moment of reading $(T+\Delta t_n)$, proportional to values X_{r1}, r = 1; 2 - to the short-term mutually correlated functions between supporting/reference and taken on the first ray/beam by signals. Remaining ray/beams when they delay relative to the first more than in 1/F, will not create on the last/latter pair of the multipliers of the noticeable voltage of frequency Δ_1. Page 157.

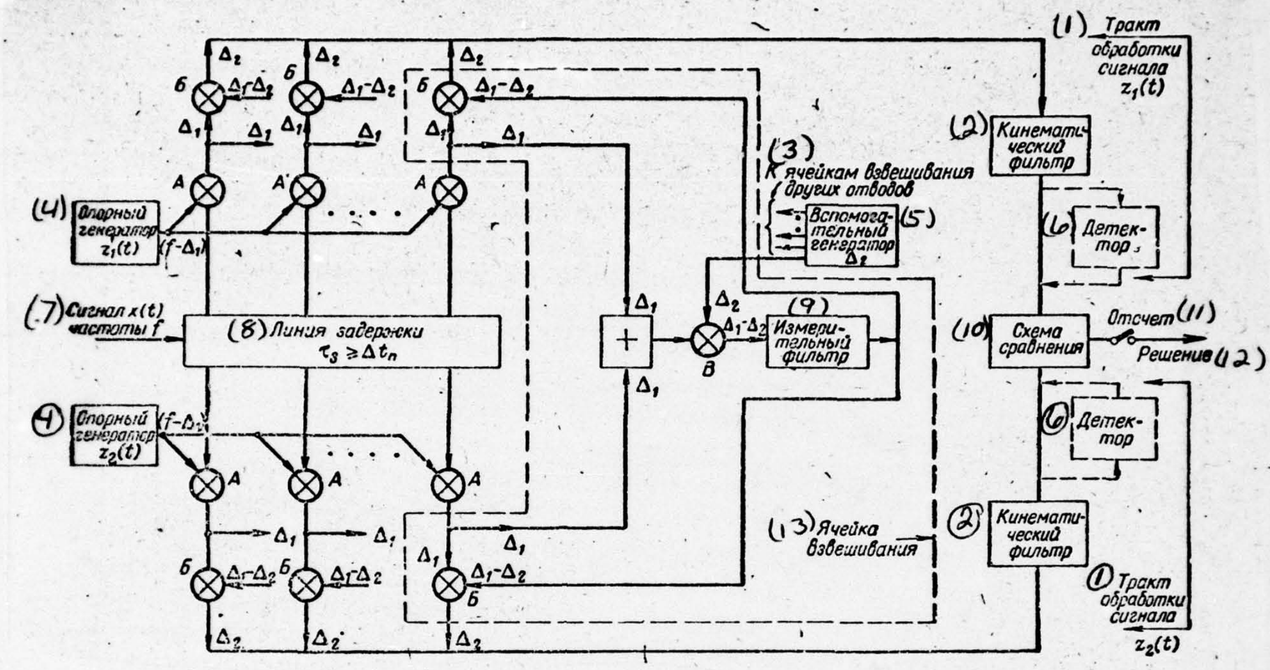
(1) Тракт обработки сигнала $z_1(t)$
(2) Кинематический фильтр
(3) К ячейкам взвешивания других отводов
(4) Опорный генератор $z_1(t)$
(5) Вспомогательный генератор Δ_2
(6) Детектор
(7) Сигнал x(t) частоты f
(8) Линия задержки $\tau_3 \geq \Delta t_n$
(9) Измерительный фильтр
(10) Схема сравнения
Отсчет (11)
Решение (12)
(4) Опорный генератор $z_2(t)$
(6) Детектор
(13) Ячейка взвешивания
(2) Кинематический фильтр
(1) Тракт обработки сигнала $z_2(t)$

Fig. 3.3.7.

Fig. 3.3.7.

Key: (1). Circuit of processing signal. (2). Kinematic filter. (3). To the cells of the weighing of other removal/outlets. (4). Supporting/reference generator. (5). Auxiliary generator. (6). Detector. (7). Signal x (t) of frequency f. (8). Delay line. (9). Measuring filter. (10). Comparison circuit. (11). Reading. (12). Solution. (13). Cell of weighing. Page 158.

However, each of them on some one of removal/outlets will turn out to be that which was synchronized with reference oscillator with accuracy not worse 1/F and it will create on the appropriate multipliers of series A the voltage of frequency Δ_1 with the instantaneous value, proportional to value for this (I-th) ray/beam. Thus, all the incoming ray/beams will be divided after the multipliers of series A.

However, the voltages in them do not satisfy another the condition of phasing. Actually, for the coherent addition of the ray/beams of the phase of the voltages of frequency Δ_1 from each of them they must be identical at the torque/moment of reading $(T+\Delta t_n)$. This means that the accuracy of the synchronization of reference

oscillator with each ray/beam must be according to condition (3.2.16) by an order less than value 1/f or, at least, not worse than 1/f. At the same time the "guaranteed" accuracy of synchronization after passage by the ray/beams of delay line because time of their time lag not multiply 1/F and randomly changes in time, composes value, although greater than 1/F however substantially exceeding 1/f. The disturbance/breakdown of synchronization can be caused also by certain inaccuracy in the installation of the removal/outlets of delay line.

Besides phasing for optimum processing multiple-pronged signal it is necessary to perform an even amplitude weighing of each of the incoming ray/beams: the contribution of I-th ray/beam to total energy of the isolatable signal must be proportional to the transmission factor of this ray/beam μ_i, i.e., is proportional to stress level, created by each of them. The voltages of the branches of processing in which the adopted ray/beams are absent and which are located under the effect only of interferences, must be to the maximum degree suppress. Of the operations of the weighing of the adoptel ray/beams and measurement of their transmission factors when using a synchronous heterodyning it is possible to combine in the weighhing equipment/device. This equipment/device of diagram in Fig. 3.3.7 consists of the separate identical cells of the weighing whose amount

is equal to the number of pairs of the removal/outlets of delay line.

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Figure 3.3.7 shows the functional diagram only of cell of weighing, which encompasses the last/latter pair of removal/outlets. This diagram is isolated in figure by dotted line. It consists of summator, multiplier (mixer) V, measuring filter and multiplier (mixer) b. The input of the summator of cell enter the voltages of frequency Δ_1 from multipliers A of the last/latter pair of the removal/outlets both of the circuit of processing signal z_1 (t) and circuit z_2 (t). This must guarantee the identical results of measurement independent of the available relationship/ratios between signals z_1 (t) and z_2 (t) during their transmission. Since the utilized signals are broadband, at the output/yield of summator we have a voltage of frequency Δ_1, proportional to value

$$x_1(t)[z_1(t-\Delta t_n)+z_2(t-\Delta t_n)],$$

where $x_1(t)=\mu_1 z_r(t-\tau_{s1})+\xi(t)$; r = 1 or 2 - the adopted only on the first ray/beam signal; $\tau_{s1}=\Delta t_n$ - the delay time in the signal on the first ray/beam. In mixer into this voltage it mixes itself with the oscillations of the auxiliary sine wave oscillator of frequency Δ_2, of common/general/total for all cells weighing. The fluctuations of the difference frequency of $\Delta_1-\Delta_2$ are supplied to narrow-band measuring filter. The latter measures the value of the correlation function between x_1 (t) and $z_1(t-\Delta t_n)$ or $z_2(t-\Delta t_n)$, proportional to transmission factor in the first ray/beam μ_1. In all cases the value of time of integration T_m measuring filter must be not less than the duration of the cell/element of useful signal T and not exceed the value, which is allow/assumed by the rate of change in the structure of multiple-pronged channel (by rate of the fluctuations of the state of the ionosphere). The value T_m is determined by the selection of the passband of measuring filter. If time of integration in measuring filter is selected by sufficiently large, then the voltage on its output/yield barely depends on fluctuating interference ξ (t) and is determined in essence by the main power of the workable ray/beam.

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Is analogous output potential of the I-th measuring filter, which

processes the appropriate synchronized ray/beam, it is determined in essence only of with a power I-th ray/beam (by value μ_i). As concerns the measuring filters, which stand in the removal/outlets, where not one of the come ray/beams did not turn out to be that which was synchronized, then output potential of such filters are virtually equal to zero.

Voltages by the frequency of $\Delta_1-\Delta_2$, proportional to value μ_i, from the output/yield of measuring filters proceed to the multipliers (mixers) of b, to which simultaneously from multipliers A will be feed/conducted the voltages by frequency Δ_1 with the instantaneous values, proportional to values X_{ri} for each I-th ray/beam. At the output/yield of each of the multipliers b is formed the voltage by frequency Δ_2, instantaneous value of which is proportional to the product of the output voltages of multiplier A and of the corresponding measuring filter, i.e., to value $\mu_i X_{ri}$. As a result of this each of the workable ray/beams turns out to be that which was suspended in accordance with its transmission factor (by power level). Voltages on the output/yield of those multipliers b, on which are absent the workable synchronized ray/beams, are virtually equal to zero.

The phasing of the workable ray/beams in diagram in Fig. 3.3.7 is realized as a result of the fact that the initial phases of the voltages of frequency Δ_2 on the output/yields of multipliers b in all removal/outlets are identical and coincide with the initial phase of auxiliary generator. Actually, let θ_i - the initial phase of output potential of I-th multiplier A, and θ_2 is a phase of the introduced frequency θ_2 [1].

FOOTNOTE [1]. Recall that the reference oscillators must be synchronized in the diagram in question with the first of the incoming ray/beams up to the initial phases in the output/yields of multipliers A it appears as a result of random change in the time lag of remaining ray/beams and inaccuracy in the installation of the removal/outlets of delay line. ENDFOOTNOTE.

Then the initial phase of voltages by the frequency of $\Delta_1-\Delta_2$, which enter from measuring filters the multipliers Б, is equal to $\theta_i-\theta_2$. The initial phases of output potentials of multipliers Б are equal to $[\theta_i-(\theta_i-\theta_2)]=\theta_2$. But θ_2 it is the constant phase of the voltage of the auxiliary generator, common/general/total for all cells of weighing. Thus, all the isolatable ray/beams are summarized in the common/general/total busbars of the circuits of processing signals z_1 (t) and z_2 (t) with

identical phases.

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Consequently, here is realized the coherent addition of the adopted ray/beams. Fold thus voltage from the busbars of the circuits of signals z_1 (t) and z_2 (t) enter the kinematic filters, tuned to a frequency Δ_2, which screen other harmonics. The instantaneous values of voltages from the output/yield of filters enter the comparator, on which at the torque/moment of reading $(T+\Delta t_n)$ on larger of them is accepted the solution to the transmission either of signal z_1 (t), or z_2 (t). It is not difficult to see that in this case conducts the coherent reception of the signals, obtained as a result of the coherent addition of ray/beams.

With processing broadband multiple-pronged signal is feasible also incoherent reception with the coherent addition of ray/beams [36]. The in principle decisive schematics of the receptors, which realize this reception, can be constructed either on multipliers or on matched filters, or with the use of a synchronous heterodyning. However, their practical realization at the present time, apparently,

most simply can be realized only in the latter case.

In the preceding/previous paragraph it was noted that when using a synchronous heterodyning the amplitude (envelope) of output potential of kinematic filter was proportional at the torque/moment of reading to value V_r - to the value envelope short-term crosscorrelation function between the adopted and reference $z_r(t)$ signals. The amplitude of the output voltage of filter can be isolated with the aid of detector. Therefore if we in diagram in Fig. 3.3.7 at the output/yield of kinematic filters include/connect envelope detectors (as this is shown in figure by dotted line), then this receptor will realize incoherent reception with the coherent addition of ray/beams. Of course, requirements for the accuracy of phasing during the coherent addition of ray/beams in it are analogous to requirements in diagram with coherent reception. At the same time with incoherent reception significantly descend requirements for the accuracy of reading of the voltages in comparator. The permissible instability of reading in this case must be by an order less than value T. For a coherent reception it must be by an order less than $1/\Delta_2$, but $1/\Delta_2 \ll T$.

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The decisive diagram of noncoherent reception with the coherent addition of ray/beams is realized in the receiver of system "Rake" (§3.5).

It is possible to construct also the diagrams of the reception of broadband signals, which realize an incoherent addition of the incoming ray/beams. Such decisive diagrams are optimum for the cases, when either due to high rate of fadings or due to the *indeterminancy*/uncertainty of the initial phase during transmission, or as a result of the limited equipment possibilities the determination of the values of initial phases (time lag Δt_i) and of the transmission factors of ray/beams μ_i is impossible. As the example, which illustrates the characteristic features of such diagrams, let us give receiver circuit with the incoherent addition of ray/beams on matched filters [36].

In §3.2 it was shown, which diffraction output potential of the filter, matched with transmitted broadband signal $z_r(t)$, has maximums for each of the adopted ray/beams, which are not superimposed one on

top of the other, if path differences between adjacent ray/beams exceed value 1/F (see Fig. 3.2.4). At the moments of time $T+\Delta t_i$ these voltages are proportional to values V_{ri} - by the envelope of short-term crosscorrelation functions between the adopted in I-th ray/beams signals and "reference" signals $z_r(t)$. For the construction of circuit with the incoherent addition of ray/beams it suffices to rectify the output voltages of filters in the circuits of processing signals z_1 (t) and z_2 (t) by squa law detectors, then to accumulate the results of detection in each circuit and tax for comparison circuit. This is realized in the binary diagram of the reception of multiple-pronged broadband signals, presented in Fig. 3.3.8. The detected output voltages of matched filters enter the summators - reservoir capacitors. Each condenser/capacitor is connected to the diagram of detector at the torque/moment (T + Δt_1) of the termination of cell/element in the first of the come ray/beams and remains that which was connected up to torque/moment $(T+\Delta t_n)$ - the reception of the latter from the incoming ray/beams. At the moment of time $(T+\Delta t_n+\varepsilon)$, where $0<\varepsilon\ll T$, condenser/capacitors are disconnected from detector diagrams and rapidly are discharged, being prepared for the arrival of the following cell/element of signal.

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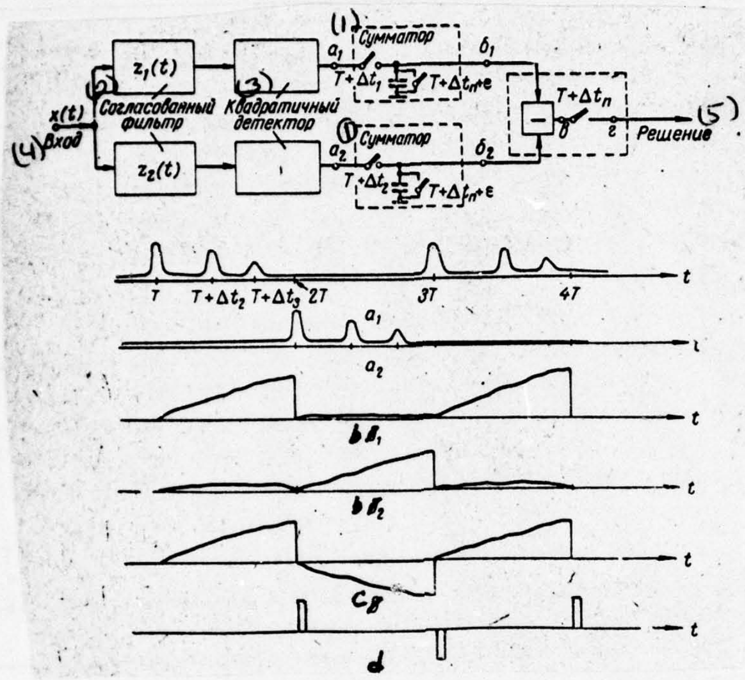
Voltages from the output/yield of condenser/capacitors proceed to the comparing subtractor, in which at torque/moment $T+\Delta t_n$ conducts the reading of voltages and in accordance with their sign is accepted the solution to the transmitted signal.

On diagram/curves are shown the voltages at the different points of diagram with the reception of the sequence of signals z_1, z_2, z_1. It is not difficult to see that the diagrams of the incoherent addition of ray/beams as compared with the diagrams of coherent addition differ in terms of simplicity. Specifically, in them considerably less stringent requirements for the synchronization receiving and transmitters.



(1) Сумматор
$z_1(t)$
a_1
$б_1$
$T+\Delta t_1$
$T+\Delta t_n+\varepsilon$
$T+\Delta t_n$
(2) Согласованный фильтр
(3) Квадратичный детектор
$x(t)$
(4) Вход
(1) Сумматор
$z_2(t)$
a_2
$б_2$
$в$
$г$
Решение (5)
$T+\Delta t_2$
$T+\Delta t_n+\varepsilon$
T
$T+\Delta t_2$
$T+\Delta t_3$
$2T$
$3T$
$4T$
t
a_1
a_2
b $б_1$
b $б_2$
c $в$
d

Fig. 3.3.8.

Fig. 3.3.8.

Key: (1). Summator. (2). Matched filter. (3). Square law detector. (4). Input. (5). Solution. Page 164.

However, the replacement of coherent addition incoherent lowers the freedom from interference of communicating system, which with high contemporary requirements for the correctness of the transmission of information can turn out to be in a number of cases unsatisfactory. Let us note also that the construction of the filters, matched with broadband complex form, represents fairly complicated technical problem.

The application/use of broadband signals as it follows from preceding/previous, makes it possible to raise the reliability of radio communication in channels with multiple-pronged propagation by means of the useful energy storage of the incoming in separate beams of signals. In this case is removed the harmful effect of selective fadings and phenomenon of the echoes, especially characteristic for short-wave radio channels.

Depending on the a priori information about the parameters of the multiple-pronged channel (time lag, the transmission factors of ray/beams) or of the possibilities of the measurement of these parameters, which in the final analysis is determined by the state of multiple-pronged channel and by contemporary technical capabilities, it is possible to construct the different diagrams of the reception of multiple-pronged signals. The manifold of such diagrams is illustrated with the aid of the diagram of Fig. 3.3.9. At present in Soviet and foreign literature appears the increasing number of works, dedicated to the investigation of the questions of effectiveness and practical realizability of such systems.

Some results according to a comparative evaluation of the freedom from interference in this chapter of the mutually correlated systems of broadband communication/connection in question and on their practical realization for short-wave channels are examined in the subsequent paragraphs.

§3.4. Freedom from interference of broadband mutually correlated systems.

Let us produce a comparative evaluation of the correctness of the transmission of the discrete information of the examined in the preceding/previous paragraphs broadband mutually correlated systems. As the measure, which characterizes the freedom from interference of such systems, let us utilize a composite probability of error with the piece-by-piece reception of the transmitted report/communication (see §2.7.), or simply the probability of error. In this case we will be restricted to the analysis of binary systems with the active pause, which include virtually the majority of the systems of military radio communication. Page 165.


(1) Решающие схемы обработки многолучевого сигнала
(2) С когерентным сложением лучей
(3) С некогерентным сложением лучей
(4) Когерентный прием
(5) Некогерентный прием
(6) На перемножителях
(7) На согласованных фильтрах
(8) С использованием синхронного гетеродинирования
(6) На перемножителях
(7) На согласованных фильтрах
(8) С использованием синхронного гетеродинирования

Fig. 3.3.9.

Fig. 3.3.9.

Key: (1). Decisive diagrams of processing multiple-pronged signal. (2). With the coherent addition of ray/beams. (3). With the incoherent addition of ray/beams. (4). Coherent reception. (5). Incoherent reception. (6). On multipliers. (7). On matched filters. (8). With use of synchronous heterodyning. Page 166.

The a priori probabilities of the transmission of signals z_1 (t) and z_2 (t) we set/assume known and equal $^1/_2$. As concerns the character of the noise, which enters from the communication channel to the input of receptor, let us consider it normal fluctuating noise with the uniform in the band of frequencies of received signals spectrum (normal white noise).

Let us pause at first at the broadband systems of single-ray reception. As it was shown in §3.2, the decisive diagrams of the receptors of these systems are realized analogous with the optimum according to Kotelnikov receivers for channels with fluctuating noise and the single-ray emission of the signal. Therefore for the mutually correlated broadband systems, in which is processed only one of the incoming ray/beams, are valid the relationship/ratios according to

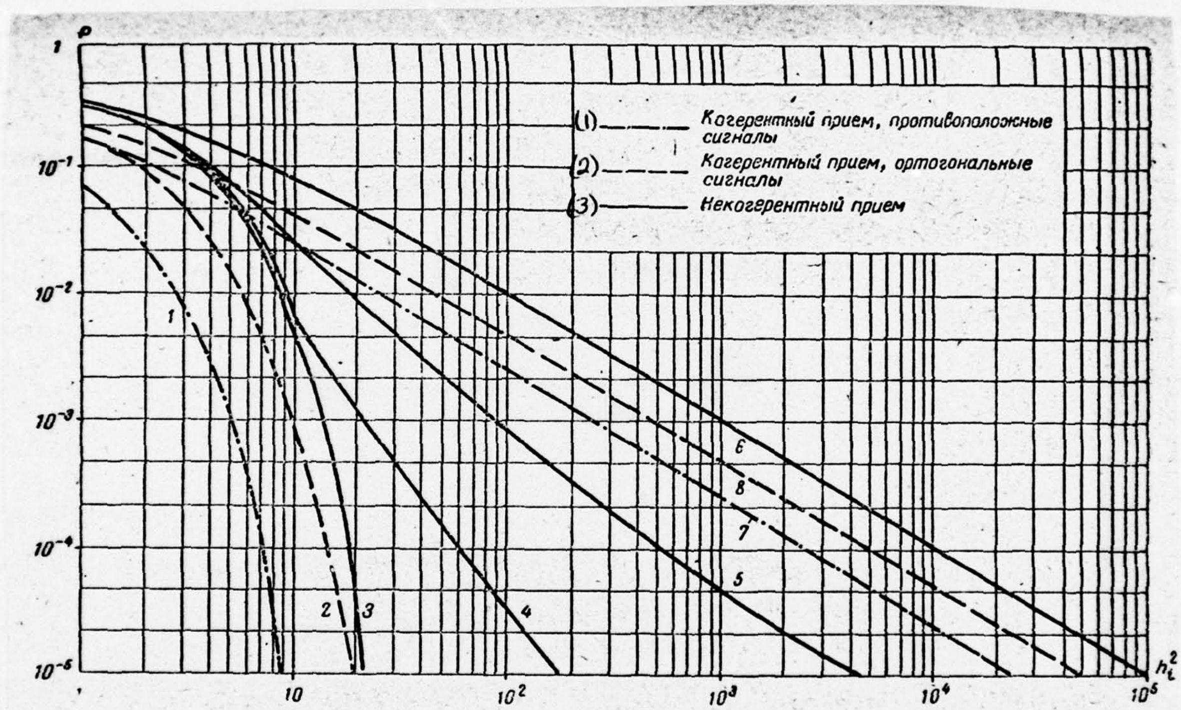
the estimate/evaluation of the freedom from interference of such optimum receivers, given in §2.7. Then in the absence signal fading in the workable ray/beam, i.e., with the constant transmission factor of this ray/beam, the probability of the error in the case of coherent and incoherent reception is determined by relationship/ratios (2.7.17), (2.7.19) and (2.7.23).

Figure 3.4.1 depicts the constructed according to these formulas dependences of the probability of the error on value $h_i^2=\frac{P_{ci}T}{\nu^2}$ - the ratio of the energy of signal in the I-th workable ray/beam to the spectral density of fluctuating interference ν^2. Curve 1 corresponds to the case of the coherent reception of the nonfading opposite signals, curve 2 - the coherent reception of orthogonal signals, curve 3 - the incoherent reception of orthogonal (in the "intensive sense") signals. As is evident from figure, the curves of the probabilities of errors monotonically decrease (freedom from interference of communicating system monotonically grow/rises) with the increase in the ratio of the energy of signal to the spectral density of fluctuating interference. In this case the knowledge and account with the reception of the initial phase of high-frequency filling of signal (coherent reception) makes it possible to obtain the higher probability of the transmission of information. On the other hand, the same probability of error is reached for coherent

reception at smaller values h_i^2, than for incoherent. Page 167.

Fig. 3.4.1.

Key: (1). —·—·— coherent reception, opposing signals. (2). --- Coherent reception, orthogonal signals. (3). ——— incoherent reception.



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At present in the transmission systems of discrete information are presented very high requirements for the correctness of reception. The probabilities of errors in this case must be less than 10^{-4}-10^{-5}. From Fig. 3.4.1 it follows that the energy loss in the case of incoherent reception and with such high requirements for the correctness of the transmission of information composes value of approximately 4 dB in comparison with the coherent reception of opposite signals and not more than 1 dB (15o/o) in comparison with

the coherent reception of orthogonal signals. However, coherent reception imposes heavy demands on the accuracy of the synchronization supporting/reference and received signals (see §3.2). With incoherent reception these requirements are significantly below. Therefore in a number of cases can turn out to be more advisable the use of an incoherent reception. In this case the decrease in the correctness of reception, caused by the nonutilization of the information about the initial phase of signal, can be compensated for by certain increase in the power of signal, which for orthogonal signals will be small.

In practice during processing I-th from the incoming ray/beams sufficiently frequently are encountered the cases, when the signal, which enters in this ray/beam, is subjected to fadings, which, however, bear common character. This character of fadings is determined by the fact that each I-th ray/beam is the beam of elementary ray/beams (microray/beams) with random amplitudes and at times time lag, much smaller than value 1/F, where F is a band of frequencies of the utilized signals.

As a result of common/general/total fadings the incoming in I-th ray/beam signal differs from that transmitted by random transmission

factor μ_i. For the analysis of the freedom from interference of the broadband transmission systems of discrete information under these conditions it is necessary to know the laws of the probability distribution of random variable μ_i. In this case in accordance with the character of its laws of distribution distinguish the Rayleigh and generalized Rayleigh (quasi-Rayleigh) fadings. Rayleigh fadings are characterized by the fact that the transmission factor has the one-dimensional probability density, which is subordinated to Rayleigh law (see §2.6).

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With the generalized Rayleigh fadings one-dimensional probability density satisfies the generalized Rayleigh distribution:

$$W(\mu)=\frac{2\mu_i}{\overline{\mu_{\phi i}^2}}\, e^{-\frac{\mu_i^2+\mu_{pi}^2}{\overline{\mu_{\phi i}^2}}}\, I_0\left(2\frac{\mu_i\mu_{pi}}{\overline{\mu_{\phi i}^2}}\right),\ \mu_i\geqslant 0.$$

The onset of the generalized Rayleigh fadings during ionospheric radiowave propagation can be explained by the fact that along with the diffuse scattering, which forms fluctuating part of the transmission factor ($\mu_{\Phi i}$), occurs the mirror reflection, which determines it regular are frequent (μ_{pi}). In certain cases the formation/education of the regular component of transmission factor can occur due to the presence of the direct/straight passage of ray/beam along the surface of the Earth into the point with time. It is obvious that Rayleigh fadings are the special case of the generalized Rayleigh fadings, when $\mu_{pi}=0$.

In the range of the average and short waves the Rayleigh and generalized Rayleigh fadings are encountered approximately equally frequently. In the range of ultra short waves during remote ionospheric and tropospheric propagaticn predominate Rayleigh fadings, and during near propagation - the generalized Rayleigh with the sharply pronounced regular component.

In the case of incoherent single-ray reception and with the generalized Rayleigh fadings the probability of the error in broadband system is equal to [36]

$$p = \frac{k^2+1}{h_{0l}^2 + 2(k^2+1)} e^{-\frac{k^2 h_{0l}^2}{h_{0l}^2 + 2(k^2+1)}}, \quad (3.4.1)$$

where $k^2 = \frac{\mu_{pl}^2}{\mu_{\phi l}^2}$ - relation regular and that which fluctuate the components of transmission factor μ_l.

Hence it follows that with the Rayleigh fadings, which are characterized by the absence of the regular component of transmission factor ($\mu_{pi}=0$, and consequently, k = 0), the probability of the error of piece-by-piece incoherent reception is determined by the relationship/ratio

$$p=\frac{1}{h_{0i}^2+2}. \quad (3.4.2)$$

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In the absence of fadings ($\mu_{\Phi i}=0,\ k\to\infty$) formula (3.4.1) passes in (2.7.23).

Figure 3.4.1 gives the dependences of the probability of error p on the mean statistical value of the ratio of the energy of signal in the workable ray/beam to the spectral density of the fluctuating interference h_{0i}^2, constructed according to (3.4.1) and (3.4.2). Curve 4 corresponds to case k^2 = 10, curve 5 - to case k^2 = 5 and curve 6 - to case k^2 = 0. Let us note that the curve 3, constructed according

to (2.7.23), corresponds to case $k^2 \to \infty$.

In the case of coherent reception and generalized Rayleigh fadings it is impossible to determine the probability of the error in the locked form. However, in the particular case of Rayleigh fadings with $h_{0l}^2 \gg 1$ it is equal to

$$p \approx \frac{1}{2\gamma^2 h_{0l}^2}. \qquad (3.4.3)$$

The dependences of the probability of error with coherent single-ray reception and Rayleigh fadings are shown in Fig. 3.4.1. Curve 7 corresponds to the application/use of opposite signals ($\gamma^2 = 2$), while curve 8 - to the case of orthogonal signals ($\gamma^2 = 1$). As can be seen from figure, the presence of common/general/total Rayleigh fadings sharply makes the freedom from interference of the single-ray reception of broadband signals worse. The probability of error in this case both in the case of incoherent and coherent reception, it proves to be to inversely proportional value h_{0l}^2 unlike channel without fadings, where dependence of p on h_{0l}^2 is close to exponential. Therefore with Rayleigh signal fading in the workable

decreases the desired values h_i^2 for obtaining small probabilities of errors. For example, already in the relation of squares regular and fluctuating component $k^2 = 5$ required for $p = 10^{-4}$ value h_{0i}^2 is equal to 500, but with $k^2 = 10$ it composes value 80. Thus, the presence of the generalized Rayleigh fadings considerably improves the conditions of reception.

Let us note also that application/use of a coherent reception and orthogonal signals in all examined above cases does not give energy gain more than 3 dB.

During processing one of the ray/beams of the incoming multiple-pronged signal, it is necessary to consider the interference of the remaining nonprocessed ray/beams. However, in the case of using broadband signals with base FT >> 1 decrease in the correctness of reception from the interference of these ray/beams can be very small.

In §3.2 it was established/installed that all ray/beams, except that which is taken, operate on receiver approximately just as supplementary fluctuating interference with the uniform spectrum of

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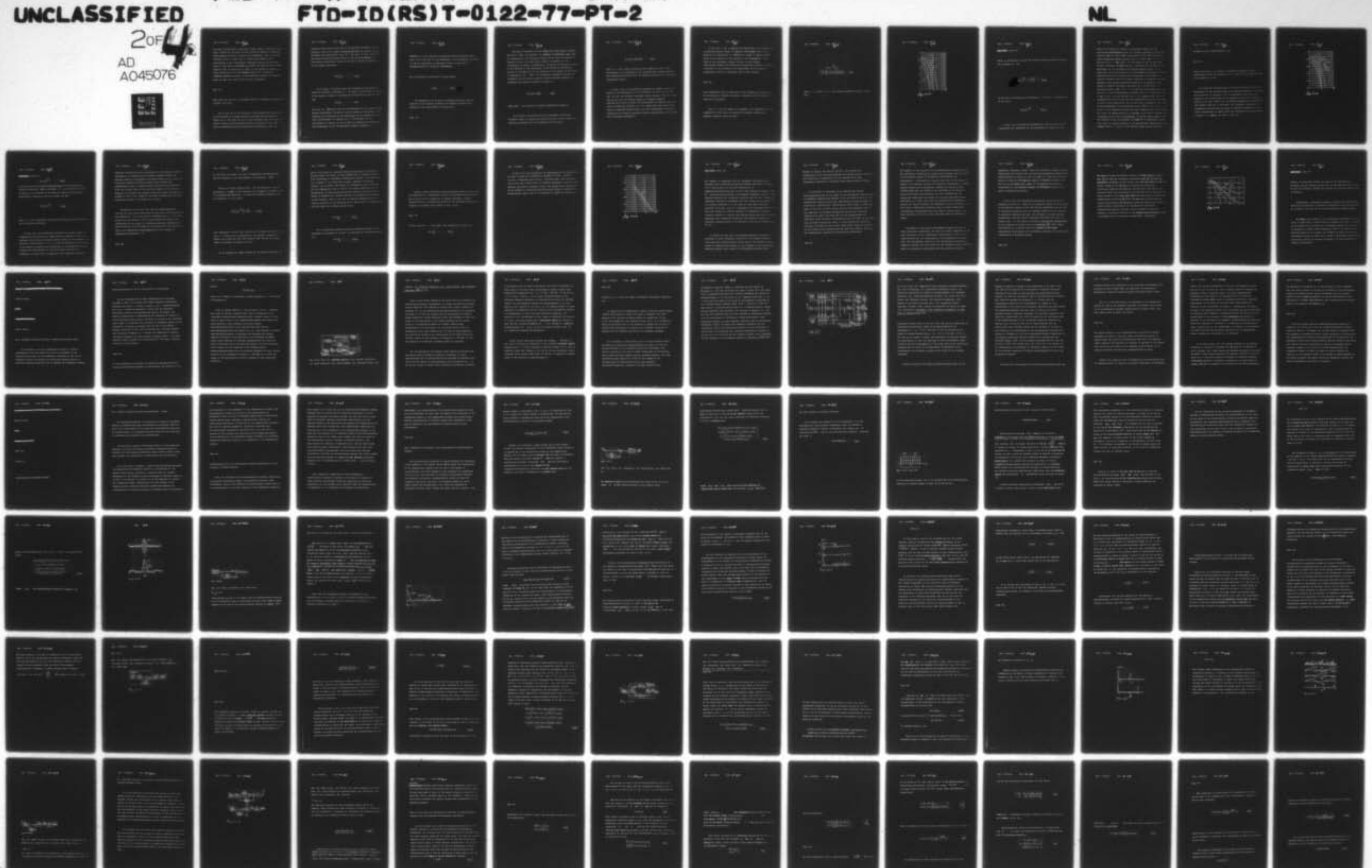

the band of frequencies F and with a power, equal to the sum of their power. Usually in the range of short waves the number of ray/beams, commensurable according to power with fundamental, does not exceed two-three. Then in order that the energy loss because of the interference of the nonprocessed ray/beams would be not more than three or four times, sufficient to provide the value of base FT with required order of magnitude h_i^2 [36]. In the absence of fadings or under conditions of the most characteristic for a single-ray reception broadband signals of quasi-Rayleigh fadings the required value h_i^2 does not exceed several dozen or hundreds.

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This value for the base of broadband system is technically completely possible (see §3.5).

Let us turn now to the analysis of the freedom from interference of the diagrams of optimum processing multiple-pronged broadband signal. As it was shown into §3.3, such receptors give off energy of entire incoming multiple-pronged signal and because of this must provide considerably higher freedom from interference, than the

diagrams, which process only one of the incoming ray/beams. Let us propose at first that signal fading in ray/beams are absent. Let us introduce into examination value $h_э^2$ - the ratio of total energy of the adopted multiple-pronged signal to the spectral density of fluctuating interference, equal to the sum of values h_i^2 from all n of the adopted ray/beams:

$$h_э^2 = \sum_{i=1}^{n} h_i^2. \tag{3.4.4}$$

If the input of receiver enter the ray/beams of approximately identical intensity, whereupon h^2 - the ratio of the energy of signal to the spectral density of fluctuating interference in each of them, then

$$h_э^2 = nh^2. \tag{3.4.5}$$

Let also T_m - effective time of the measurement of the parameters of medium (transmission factors and initial phases of ray/beams), that considers the prehistory of the measurement of the parameters to the given torque/moment of reading, but T - the duration of the cell/element of signal, at end of which is accepted the solution to the transmission of the corresponding version of signal [1].

FOOTNOTE [1]. In the diagram of processing multiple-pronged signal (Fig. 3.3.7) the time of the measurement of the parameters of medium is time of integration of measuring filter in each cell of equipment/device of weighing. ENDFOOTNOTE.

Let us designate the relation of these values

$$\delta = \frac{T_m}{T}. \quad (3.4.6)$$

The probability of the error in diagrams with the coherent addition of the incoming ray/beams and coherent reception for arbitrary n was investigated in report [45].

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Virtually interesting are the cases $\delta\to\infty$ $(T_m\gg T)$ and $\delta\to 1(T_m \doteq T)$. The case $\delta \longrightarrow \infty$ corresponds, for example, to situation, when time of integration in the measuring filters of the cells of weighing in diagram in Fig. 3.3.7 many times exceeds the duration of the cell/element of signal. Consequently, the effect of fluctuating noise on the accuracy of the measurement of the values of transmission factors in ray/beams and to the accuracy of the phasing of ray/beams is negligible. For $\delta \longrightarrow \infty$ the orthogonal transmitted signals the probability of the error in system with the coherent addition of ray/beams and coherent reception is equal to

$$p^{(n)}=\frac{1}{2}[1-\Phi(h_0)], \qquad (3.4.7)$$

where $\Phi(x)$ - the function of Kramer (probability integral).

If the input of receiver enter the ray/beams of identical intensity, then, by taking into account (3.4.5), we will obtain the following expression for the probability of the error:

$$p^{(n)} = \frac{1}{2}\left[1 - \Phi\left(\sqrt{n}h\right)\right]. \qquad (3.4.8)$$

With n = 1, this formula coincides with formula (2.7.19), which corresponds to the reception of one ray/beam with a priori known by the transmission factor and by the initial phase with the orthogonal transmitted signals.

In Fig. 3.4.2 are constructed according to formula (3.4.8) the curves of the dependences of the probability of error p on h^2 for one, two and four workable ray/beams (dotted curves). From the figure one can see that the energy storage of ray/beams proviles substantially higher freedom from interference as compared with the case of single-ray reception (n = 1). Energy gain when using total energy of multiple-pronged signal with the approximately identical intensities of separate ray/beams directly proportional to the amount of the incoming ray/beams n.

In the case $\delta \rightarrow 1$, time of the measurement of the parameters of multiple-pronged signal T_m becomes commensurable with to the duration of cell/element T. A decrease in value T_m can be caused either by an increase in the velocity of the fluctuations of the state of the ionosphere (signal fading), or with design considerations during the production of measuring filters it leads to a decrease in the accuracy of the measurement of the values of transmission factors in ray/beams and in their phasing.

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This unavoidably must be reflected in the freedom from interference of the system of coherent reception in question with the coherent addition of ray/beams.

With $\delta = 1$ and the number of ray/beams $n \geqslant 2$ expression for the probability of the error in system with coherent addition and coherent reception takes the form

$$p^{(n)} = e^{-\frac{h_э^2}{2}}\left[\frac{1}{2} + \sum_{m=1}^{n-1} \frac{h_э^{2m}}{m!\,2^m} \sum_{j=m}^{n-1} \frac{(n-1+j)!\,2^{-j-n}}{(n-1+m)!\,(j-m)!}\right], \qquad (3.4.9)$$

where k ! = 1•2•3• ... •k - the factorial positive integer number [20].

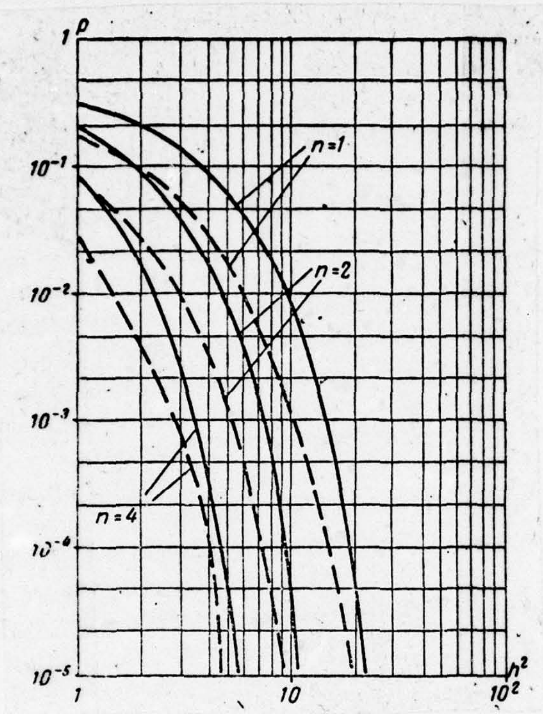

Fig. 3.4.2.

[redacted]. Page 175.

Hence, in particular, we have the following formula during processing two ray/beams (n = 2):

$$p^{(2)}=\frac{1}{2}e^{-\frac{h_{э}^{2}}{2}}\cdot\left(1+\frac{h_{э}^{2}}{8}\right). \qquad (3.4.10)$$

In the case of processing one ray/beam (n = 1) and δ = 1 probability of the error

$$p^{(1)}=\frac{1}{2}e^{-\frac{h_{э}^{2}}{2}}. \qquad (3.4.11)$$

In Fig. 3.4.3 according to formulas (3.4.10) and (3.4.11) are constructed the dependences of the probability of error p on the

ratio of the energy of signal to the spectral density of the fluctuating interference h^2 for the incoming ray/beams of identical intensity and $\delta=1$ $(T_m=T)$. For a comparison in this same figure are given probability curves of error for n = 2 and $\delta \rightarrow \infty$; δ = 2.5, and also for n = 1 and $\delta\rightarrow\infty$; δ = 0.25 [45]. As can be seen from figure, a decrease in value δ, i.e., a decrease in the time of the measurement of the parameters of multiple-pronged signal T_m in comparison with the duration of the cell/element of signal T, causes deterioration in the freedom from interference of the system of coherent reception with the coherent addition of ray/beams. Moreover decrease δ more sharply manifests itself with an increase in the number of workable ray/beams. Actually, for $p \geqslant 10^{-4}$ with n = 1 transition from $\delta \rightarrow \infty$ to δ = 1 is connected with energy loss in 1 dB, i.e., is very insignificant. However, with n = 2 transition from $\delta \rightarrow \infty$ to δ = 2.5 and further to δ = 1 is connected for the same probabilities of the errors with an increase in the energy loss into 0.6 and 1.6 dB respectively. Let us note that a decrease in the time of the measurement of the parameters from $\delta \rightarrow \infty$ to δ = 1 at n = 5 and n = 10 is connected already with very essential gain on the order of 5 and 7 dB respectively [45]. From Fig. 3.4.3 also it follows that a decrease in the time of measurement T_m to the values smaller than the duration of the cell/element of signal T, is inexpedient, since this leads to a sharp decrease in the freedom from interference. For example, with n = 1 and δ = 0.25 (dotted curve) energy loss will

increase to value approximately 6 dB.

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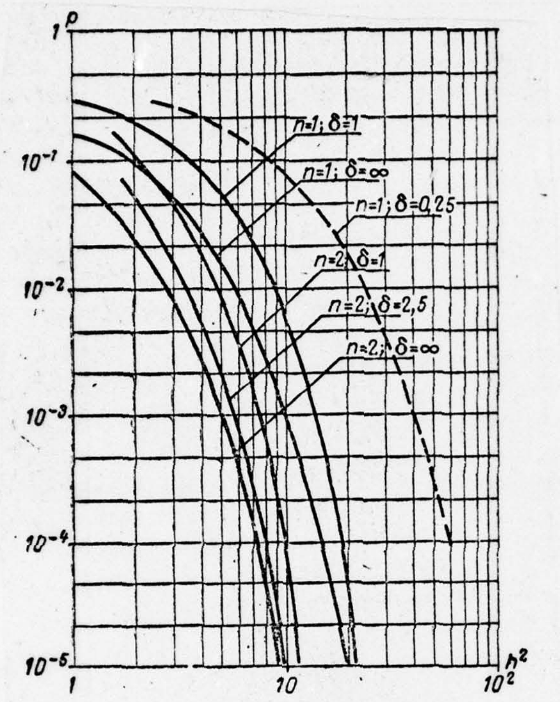
With an increase in the amount of workable ray/beams it becomes inadmissibly large. For example, at $n = 5$ and $\delta = 0.25$ value of loss will increase to 10 dB.

The expressed considerations are the very essential during the development of the measuring filters of equipment/devices of weighing in the systems in question. This, in particular, was taken into account in the construction of the measuring device of a broadband system of the type "Rake". For a broadband communicating system with the coherent addition of ray/beams, but incoherent reception, the probability of error with any number of workable ray/beams n and of time of measurement T_m, much larger the duration of the cell/element of signal T ($\delta \rightarrow \infty$), is equal to [36, 45]



P
1
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
1
10
10^2
n²
n=1; δ=1
n=1; δ=∞
n=1; δ=0,25
n=2; δ=1
n=2; δ=2,5
n=2; δ=∞

Fig. 3.4.3.

[redacted]. Page 177.

$$p^{(n)}=\frac{1}{2}e^{-\frac{h_{э}^{2}}{2}}. \qquad (3.4.12)$$

In the case of the arrival of the ray/beams of the approximately identical intensity, when $h_{э}^{2}=nh^{2}$, where h^2 - the ratio of the energy of signal in one ray/beam to the spectral density of fluctuating interference, expression (3.4.12) assumes the form

$$p^{(n)}=\frac{1}{2}\cdot e^{-\frac{nh^{2}}{2}}. \qquad (3.4.13)$$

With n = 1, this expression coincides with formula (2.7.23), obtained for a single-ray reception.

In Fig. 3.4.2 are constructed according to (3.4.13) curves the probabilities of the error in system with the coherent addition of ray/beams and incoherent reception (solid lines). As the parameter of curves are taken values n = 1, 2, 4. From the figure one can see that in the system in question the accumulation of ray/beams also raises freedom from interference. In this case energy gain for the probability of error 10^{-4} in comparison with single-ray incoherent

reception increases directly proportional to the number of workable ray/beams n. In comparison with coherent reception during the coherent addition of ray/beams (the dotted curves of Fig. 3.4.2) the incoherent reception provides somewhat lower freedom from interference, since is connected with the loss of the information about the initial phase of the resulting signal in the decisive schematic of receptor (for example, after the output of detectors in diagram in Fig. 3.3.7). However, the value of energy loss is small and does not exceed 1 dB (15o/o) for $p \geqslant 10^{-4}$.

Interesting to note that fact that of relationship/ratio (3.4.12) and (3.4.13) are valid also in the case of the reception of one ray/beam (n = 1) at any value δ, including with $\delta = 1$. In other words, with the reception only of one ray/beam the freedom from interference of the system of incoherent reception, determined by these formulas, does not depend on the selection of the duration of time of the measurement of the parameters of the channel of communication/connection T_m,

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In this case, of course, the time of measurement must be not less than the duration of the cell/element of the signal: $T_m > T$.

Further, in other limiting case - for the duration of time of measurement T_m, equal to the duration of the element of signal T, and to processing two ray/beams - we have the following expression for the probability of the error:

$$p^{(2)} = \frac{1}{2} e^{-\frac{h_э^2}{2}} \left(1 + \frac{h_э^2}{4}\right). \qquad (3.4.14)$$

From comparison (3.4.14) with (3.4.10) it is evident that with n = 2 and δ = 1 freedom from interference of the coherent reception is higher than incoherent one. This remains valid and for the larger number of workable ray/beams and δ ≫ 1.

In the presence of common fadings in the adopted ray/beams the

use of total energy of multiple-pronged signal makes it possible to largely remove the effect of these fadings and to increase freedom from interference of broadband communicating system. Let us show this in an example of the double-beam signal, subjected to Rayleigh fadings. Let us assume that fadings in ray/beams are not correlated and the intensity of each of them on the average identical. Let us designate by h_0^2 the average statistical value of the ratio of the energy of signal to the spectral density of fluctuating interference in each ray/beam. Then in the case of coherent reception with the coherent addition of two ray/beams and at values $h^2_0 \gg 1$ probability of error takes the following simple form:

$$p^{(2)} \approx \frac{3}{4h_0^4}. \qquad (3.4.15)$$

For an incoherent reception with the coherent addition of two ray/beams the probability of error is expressed by relationship/ratio [1]

$$p^{(2)} \approx \frac{2}{h_0^4}. \qquad (3.4.16)$$

Finally, during processing multiple-pronged signal in cases when the measurement of transmission factors in ray/beams hinder/hampered due to high speed of fadings or is absent according to design considerations, is feasible the method of the incoherent addition of signals in ray/beams (see diagram in Fig. 3.3.8).

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In this case for n = 2 and $h_0^2 \gg 1$ the probability of the error

$$p^{(2)} \approx \frac{3}{h_0^4}. \qquad (3.4.17)$$

In Fig. 3.4.4 are constructed the dependences of the probability of error on value h^2_0 according to formulas (3.4.15) - dotted, (3.4.16) - continuous and (3.4.17) - dot-dash line. From the figure one can see that already with two workable ray/beams, in spite of the Rayleigh character of fadings in them, the freedom from interference of all systems of processing multiple-pronged signal in question is substantially higher than in the diagrams of single-ray reception.

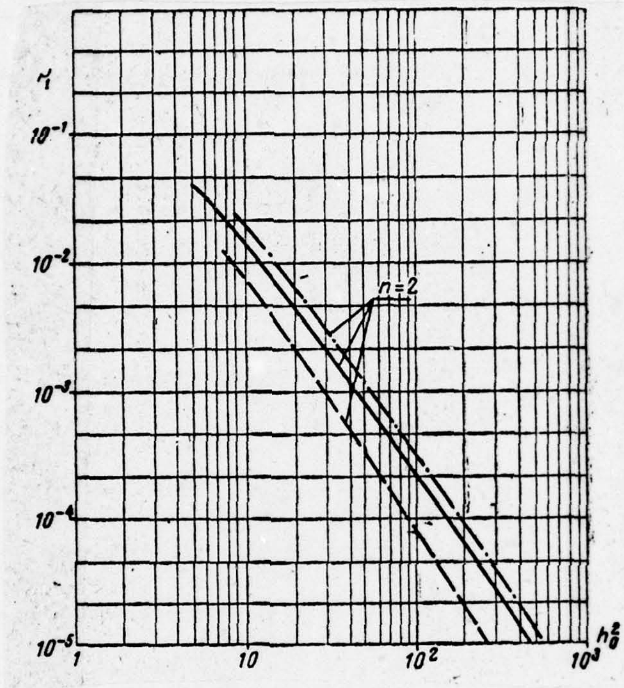

Fig. 3.4.4.

~~Fig. 3.4.4~~. Page 180.

For example, in comparison with the incoherent reception of one ray/beam under conditions of Rayleigh fadings (see curve 6 in Fig. 3.4.1) energy gains for the high authenticity of communication/connection with $p \gg 10^{-4}$ they compose the values: 20.6 dB - during coherent addition with coherent reception; 18.5 dB - with coherent addition with incoherent reception and 17.6 dB - during the incoherent addition of ray/beams. With an increase in the amount of the workable ray/beams n of the value of these gains even more grow/rise. From Fig. 3.4.4 it is possible to be convinced also of the fact that during the coherent addition of ray/beams the coherent reception provides in channel with fadings energy gain into 2.1 dB in comparison with incoherent reception, and incoherent addition of ray/beams insignificantly it is inferior to the latter, playing back 0.9 dB.

It should be noted that the Rayleigh character of fadings in ray/beams is least favorable, since out of all actually possible cases precisely Rayleigh signal fading lead to the greatest energy losses in communicating systems. In the presence of the generalized Rayleigh fadings, which occupy the intermediate position among

absence of fadings and Rayleigh fadings, the freedom from interference of the indicated diagrams of multiple-pronged reception will grow/rise in comparison with the examined above case, approaching conditions of reception under channel without fadings.

In conclusion of paragraph let us examine some special feature/peculiarities of the effect of the concentrated interferences on broadband communicating systems. As it was noted in §2.1, in the communication channels along with fluctuating are widely common the concentrated interferences. Specifically, their amount is especially great in the range of short waves, what is the consequence both of the conditions of radiowave propagation and all the increasing charging this range. The large part of such interferences has relatively small intensity. Store/adding up at the input of receptor, they is formed fluctuating noise, in their statistical properties virtually not differing from normal white noise. Under the influence of such concentrated interferences are valid all results, given for the communication channels with fluctuating interference.

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The presence of the totality of the concentrated interferences with the statistical properties of normal noise increases only the value of the spectral density of fluctuating interference, and consequently, it decreases the value of the ratio of the energy of signal to the spectral density of fluctuating interference, lowering thereby freedom from interference. However, the effect of such interferences does not depend on the structures of the utilized signals and equally detrimentally both for broadband and for narrow-band communicating systems. At the same time under actual conditions, as a rule, some from the concentrated interferences sharply are isolated against the background of their totality, i.e., have power, considerably exceeding power of other concentrated interferences commensurable with the power of useful signal, but their spectrum in full or in part coincides with the spectrum of signal.

The effect of such single interferences depends not only on their statistical properties, but also on a mutual difference in the form (structure) of the concentrated interference on the form of those utilized for the transmission of the information of signals [30]. The overwhelming majority of the concentrated interferences appears at present from the signals of the extraneous radio stations, which are sinusoidal oscillations with the changing parameters

(amplitude, frequency, phase). Under the influence of single harmonic interference the freedom from interference of broadband communicating system depends no longer not only on value h_s^2 - the ratio of total energy of signal to the spectral density of filtration noise, but also on the parameter h_π^2/FT, where h_π^2 is ratio of the energy of interference to the spectral density of fluctuating noise, FT is a base of communicating system.

In Fig. 3.4.5 are constructed probability curves of error p depending on value h^2 - ratio of the energy of signal in one ray/beam to the spectral density of the fluctuating interference for the case of incoherent reception with the coherent addition of two ray/beams of identical intensity and under the influence of single harmonic interference. Unbroken curves correspond that which not fade, and broken lines - to the fading according to Rayleigh law interference. As the parameter of curves serve the relations $\frac{h_\pi^2}{FT}=0;\ 0{,}5;\ 1.$ From curve/graphs it is evident that the presence of the single concentrated interference can considerably influence the freedom from interference of broadband system.

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The effect of this interference begins to already manifest itself with $\frac{h_n^2}{FT} \geqslant 0{,}5$. For the nonfading interference with $\frac{h_n^2}{FT}=0{,}5$ and the probability of error $p \geqslant 10^{-4}$ energy loss does not exceed 2-3 dB. The further growth h_n^2/FT, and also the presence of fadings of interference lead to its increase. Specifically, with fadings the indicated above loss grow/rises by 2-3 more dB. Let us note that the energy lossses of the same order begin to appear also in narrow-band communicating system for $\frac{h_n^2}{FT} \geqslant 0{,}5 \div 1$. At the same time from this an example escape/ensues the essential difference between the effect of single harmonic interference on broadband and narrow-band communicating systems. Actually, single harmonic interference produces a noticeable decrease in the freedom from interference both in broadband and in narrow-band systems with $h_n^2=(0{,}5 \div 1)FT$.

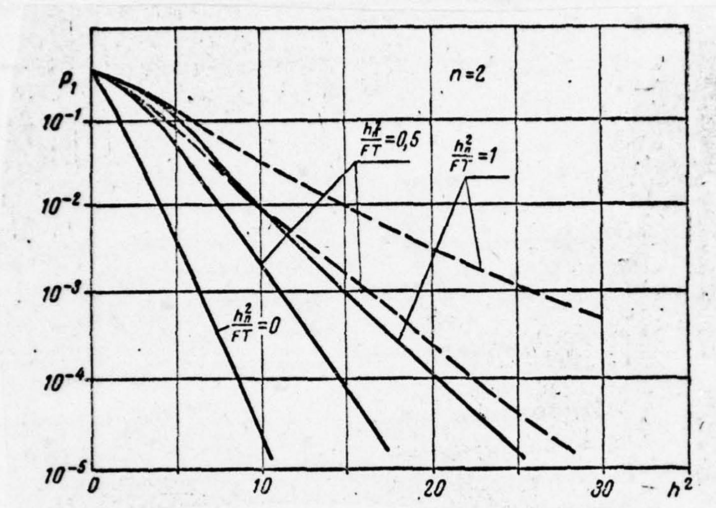
P_1
$n=2$
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
$\frac{h_n^2}{FT}=0,5$
$\frac{h_n^2}{FT}=1$
$\frac{h_n^2}{FT}=0$
0
10
20
30
h^2

Fig. 3.4.5.

[redacted]. Page 183.

however in narrow-band system the value of the utilized base virtually does not exceed several ones. At the same time of the value of base for broadband signals they compose the value of several dozen or hundreds.

Consequently, in broadband system is provided the considerably better/best suppression of harmonic interference. From this viewpoint an increase in the base of the utilized signals is highly useful.

Of course, most intense of the concentrated interferences with value h_n^2 **the** order of several dozen or hundreds can cause a decrease in the freedom from interference of broadband system. In this case it is necessary to ensure their suppression prior to the input of the decisive schematic of receptor, for example, by means of rejection sufficiently narrow in order not to cause a considerable decrease in the energy, section of the band of signal F, in which operates the harmonic interference.

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Pages 183-217.

§3.5. Broadband mutually correlated communicating system "Rake"

The developed in the USA transmission system of a discrete information of the type "Rake" can serve as an example of the practical realization of the fundamental principles of the use of broadband signals for dealing with multi-beam characteristics in shortwave communications [44]. Let us examine the fundamental special

feature/peculiarities of the construction of this system.

For the transmission of binary information from telegraph transmitter (Fig. 3.5.1) emits into space frequency- manipulated broadband with band $F = 10$ kHz of signal z_1 (t) - "premise/impulse" and z_2 (t) - "pause" by duration T, also, with the frequency shift/shear between them approximately 182 Hz. In receptor the received signals must be reconstructed for obtaining the energy of entire multiple-pronged signal. Therefore to utilize as any wide-band noise signal is impossible. The formation of broadband signals in the transmitter of system is realized with the aid of the special diagram, encircled in Fig. 3.5.1 by dotted line. Steering impulses with repetition frequency $f_0 = 120$ kHz, the stabilized by crystal oscillator frequency f_0, they proceed to special device - the shift register, which consists of the dialing/set of 10 Origger circuits, included by code feedback.

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At the output/yield of register is formed the periodically being repeated pseudorandom sequence of video pulses. The duration of its

period

$$T_0=(2^m-1)t_0,$$

where m is a number of flip-flops in shift register; t_0 - the period of frequency f_0.

Since in system "Rake" f_0 = 120 kHz and m = 10, T_0 = 8.525 ms. This value of period is selected more than the duration of the ionospheric multiple-beam characteristics, component usually in the range of short waves 3-5 ms. Pseudorandom sequence then enters the pulse generator, which develops very short pointed momentum/impulse/pulses, that develops very short pointed momentum/impulse/pulses by duration 0.3 μs. This character of these momentum/impulse/pulses is necessary for obtaining the largest possible uniformity of the spectrum of the "carrying" broadband signal whose formation is realized subsequently by the first and second band-passs filter and amplitude limiter. The first band-pass filter passes from the spectrum of the incoming pointed momentum/impulse/pulses the only harmonic components in band ±5 kHz relative to the frequency of tuning f_1 = 455 kHz. As a result the signal at the output/yield of filter acquires in time noise-like character.

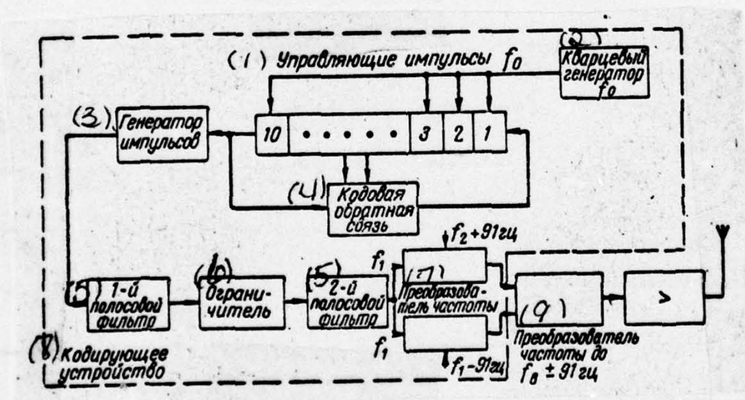

Fig. 3.5.1. Key: (1). **Steering impulses.** (2). Crystal oscillator. (3). Pulse generator. (4). Code feedback. (5). band-pass filter. (6).

Limiter. (7). Frequency converter. (8). Coding device. (9). Frequency converter to. Page 185.

Since in the output voltage of the first filter are possible the sufficiently outbursts (overshoots), it enters the bilateral limiter through amplitude. The spectrum of signal after limiter somewhat is expanded. For the preservation/retention/maintaining of the spectrum of signal F = 10 kHz the voltage from the output/yield of limiter passes through the second band-pass filter with passband 10 kHz. As a result of the fluctuation of the output potential of the second filter to a considerable degree are limited, which brings subsequently to the better/best use of energy service lives of transmitter by the more effective use of tubes of power amplifier. The spectrum of signal at the output/yield of the second filter is limited in band ±5 kHz relative to frequency f_1 = 455 kHz. On this the formation of "carrying" broadband signal it concludes.

For the realization of frequency shift keying the voltage from the second filter proceeds to frequency converters, to which simultaneously will be feed/conducted the voltages by frequency (f_2 + 91 Hz) for the signal of premise/impulse and by frequency (f_2 - 91 Hz) for the signal of pause. These last/latter voltages are switched

in accordance with the work of telegraph. The value of frequency is taken equal to 155 kHz. After transformation output voltages have respectively frequencies ($f_1 - f_2$ - 91 Hz) = 300 kHz - 91 Hz and ($f_1 - f_2$ + 91 Hz) = 300 kHz + 91 Hz. These voltages proceed to the following frequency converter, at output/yield of which are obtained finally the high-frequency signals of premise/impulse z_1 (t) and pauses z_2 (t). signals z_1 (t) and z_2 (t) are passed through the power amplifier of transmitter and are emitted by the transmitting antenna to correspondent. In time they have noise-like character. The duration of the cell/element of each signal is identical and equal to T = 22 ms, but their spectra are concentrated within the limits of band ±5 kHz relative to frequencies f_B+91 Hz for the signal of premise/impulse and f_B-91 Hz for the signal of pause. Here f_B is a operating frequency of transmitter.

During system tests were utilized the values f_B=8, 12, 17 MHz. Since the bands of frequencies of the utilized broadband signals compose value F = 10 kHz, the base of system is equal to FT = 220. It should be noted that signals z_1 (t) and z_2 (t) are orthogonal, since frequency shift between them, equal to 182 Hz, to multiple of inverse value of the duration of premise/impulse T.

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Actually, at T = 22 ms the value of frequency shift/shear comprises 4/T.

In passing by the communication channel, multiple-pronged signal together with noises it enters the input of receptor. The block diagram of the receiver of system "Rake" is represented in Fig. 3.5.2. From comparison with diagram in Fig. 3.3.7, it is possible to make the conclusion that in the receptor of system with the aid of synchronous heterodyning is realized the incoherent reception of signals with the coherent addition of all incoming ray/beams.

Let us examine in more detail, as in the given diagram occurs recording the signals of premise/impulse and pause. For this the diagram switches on two identical circuits of processing. Each circuit consists of the line of delay, dialing/set of correlators, which weighs and which phases devices, envelope detector. The taken multiple-pronged signal as a result of the passage of the cascade/stages of amplification along high frequency (ultrahigh-frequency), conversion and amplification along

intermediate frequency (UPCh) is converted into the signal of frequency f_1 = 455 kHz and enters the input of delay line. The delay line is common/general/total for circuits processing the signals of premise/impulse z_1 (t) and pause z_2 (t). Complete delay time in it τ = 3 ms. Since the width of the field of the powerful correlation of the utilized signals composes value $\pm 1/F = \pm 100$ μs, in line for each circuit is provided series from τF = 30 removal/outlets, placed through identical time intervals 1/F. Voltage by frequency f_1 = 455 kHz from the removal/outlets of line proceeds to multipliers (mixers) A of the dialing/set of correlators. Besides multiplier A each correlator switches on even reference oscillator and the integrating (kinematic) filter. Moreover both reference oscillator and the integrating filter are common/general/total for all correlators of each circuit of processing. The reference oscillators of signals z_1 (t) and z_2 (t) in receiver are the devices, similar to the generator of the formation of the broadband signals of transmitter. Page 187.

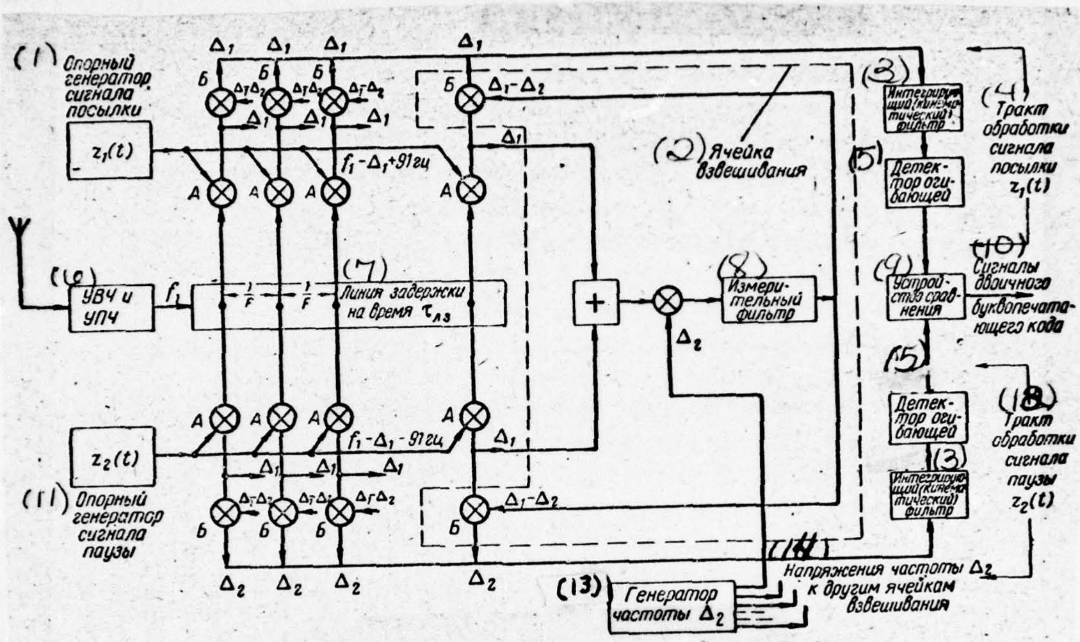
(1) Опорный генератор сигнала посылки
$z_1(t)$
(2) Ячейка взвешивания
(3) Интегрирующий (кинематический) фильтр
(4) Тракт обработки сигнала посылки $z_1(t)$
(5) Детектор огибающей
(6) УВЧ и УПЧ
f_1
(7) Линия задержки на время $\tau_{лз}$
(8) Измерительный фильтр
(9) Устройство сравнения
(10) Сигналы двоичного буквопечатающего кода
(11) Опорный генератор сигнала паузы
$z_2(t)$
(12) Тракт обработки сигнала паузы $z_2(t)$
(13) Генератор частоты Δ_2
(14) Напряжения частоты Δ_2 к другим ячейкам взвешивания
$f_1-\Delta_1+91$ гц
$f_1-\Delta_1-91$ гц
Δ_1
Δ_2
$\Delta_1-\Delta_2$
А
Б

Fig. 3.5.2.
[ГЦ = Hz]

Fig. 3.5.2. Key: (1). Supporting signal generator of premise/impulse. (2). Cell of weighing. (3). Integrating (kinematic) filter. (4). Circuit of processing the signal of premise/impulse. (5). Envelope detectors. (6). Ultrahigh-frequency and UPCh. (7). Delay line for a period. (8). Measuring filter. (9). Device of comparison. (10). Signals of the binary printing code. (11). Supporting signal generator of pause. (12). Circuit of processing the signal of pause. (13). Generator of frequency. (14). Voltages of frequency to other cells of weighing. Page 188.

Difference consists only of the fact that here for the realization of synchronous hyterodyning occurs the formation/education of frequencies $f_1 - \Delta_1 + 91$ Hz = 435 kHz + 91 Hz in the case of the signal of premise/impulse and $f_1 - \Delta_1 - 91$ Hz = 435 kHz - 91 Hz in the case of the signal of pause. The generators of reference signals are synchronized so that the beginning of their cell/element would coincide with the torque/moments of the time, when on the last/latter removal/outlet of delay line render/shows the beginning of the cell/element of the signal, accepted on the first of the incoming ray/beams.

Further processing the adopted multiple-pronged signal in the

diagram in question is based on the measurement of the short-term (for time of the duration of cell/element T) mutually correlated function of that taken on each ray/beam and supporting/reference signals. So, from the last/latter pair of the multipliers of series A will arise the voltages of frequency Δ_1 = 20 kHz, with the instantaneous values at the torque/moment of reading ($T + \tau$), proportional to the value of the mutually correlated function between the supporting/reference and that which is taken on the first ray/beam signals. Remaining ray/beams, if they delay relative to the first for a period more 1/F, will not create on the last/latter pair of the removal/outlets of the noticeable voltage of frequency Δ_1. However, each of them on some one of removal/outlets will be synchronized with reference signal with an accuracy to 1/F and it will create on the output/yield of the corresponding multiplier the voltage of frequency Δ_1 with the instantaneous value, proportional to the mutually correlated function between this ray/beam and the reference signal. Thus, all the adopted ray/beams with the time lag of each of them, that exceed value 1/F, turns out to be divided at the output/yield of multipliers A. Thereby is removed the ill effect of the selective character of fadings and phenomenon of echo on the reception of signals.

For using the total energy of entire multiple-pronged signal and

increase thereby the correctness of the reception of information the receptor of system "Rake" makes an additional two operations, substantially important during processing multiple-pronged signal.

This is, in the first place, the operation of the weighing of signals on each of the incoming ray/beams: the greater signal level in the corresponding ray/beam (removal/outlet of delay line), that more powerful must be used this signal.

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The output voltages of the removal/outlets, in which the adopted ray/beams are absent or are very small and which are located in essence under the action of interferences, must be to the maximum degree suppress. The operation of weighing is realized in this system with the aid of the device of the weighing, which consists of the separate identical for each pair removal/outlets of cells.

Figure 3.5.2 shows the cell of weighing for the last/latter pair of removal/outlets. It consists of summator, multiplier and measuring

filter. The operating principle of this cell was examined in §3.2. Here let us let us point out that in system "Rake" the passband of measuring filter is selected order 1 Hz, which approximately corresponds to the speed of the fluctuations of the state of ionospheric channel. Then time of integration of filter, inversely proportional to the value of its band, many times exceeds the duration of the cell/element of signal T. Consequently, the output voltage of the measuring filter of the frequency $(\Delta_1 - \Delta_2)$ of each cell of weighing barely depends on fluctuating noises and turns out to be proportional to the intensity of the corresponding ray/beam. It proceeds to the multipliers of series *Б*, at output/yield of which are formed the voltages of frequency Δ_2 = 9 kHz with the instantaneous value, proportional at the torque/moment of reading to the product of the short-term function of mutual correlation by the appropriate weight coefficient, determined by ray intensity.

In the second place, for the coherent addition of all adopted ray/beams in receiver is realized the operation of the phasing of their voltages. The "rough" phasing of separate ray/beams is realized by means of their synchronization with accuracy 1/F with the aid of delay line. A "precise" phasing of voltages in ray/beams conducts electrically because of frequency conversion on the multipliers of series *Б* and of the presence of the single for all cells weighing of

the generator of frequency Δ_2 = 9 kHz. Actually, if one considers that the initial phases of voltages are converted in the multipliers (mixers) of removal/outlets just as frequency, then the initial phases of the voltages of frequency Δ_2 on the output/yield of all multipliers of series Б are identical they coincide with the initial phase of the generator of frequency Δ_2.

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Fold thus voltage from the common/general/total busbars of signals z_1 (t) and z_2 (t) enter the integrating kinematic filters, which are tuned to a frequency Δ_2, and then they are introduced into the diagrams of envelope detectors. Output potentials of detectors at the torque/moment of reading are proportional to the values of the envelopes of short-term crosscorrelation function between those which were folded coherently kogerentno by signals in ray/beams and by reference signals z_1 (t) and z_2 (t). These voltages enter the comparison circuit, in which on larger of them is accepted the solution to the reception either of the signal of premise/impulse or the signal of pause. The output voltage of comparator controls the work of telegraph at the point of reception.

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Chapter 4.

AUTOCORRELATION BROADBAND SYSTEMS.

§4.1. General information about autocorrelation systems.

The isolation/liberation of signals out of noises in broadband systems is possible also when using methods of reception, which are based on the measurement of the short-term autocorrelation function of received signal. The broadband systems, which use this principle, is conventionally designated as autocorrelation.

The distinctive special feature/peculiarity of autocorrelation systems is the fact that during correlation processing signal in receiver the role supporting/reference plays received signal itself, more precise, the adopted sum of useful signal and interferences.

As it was shown in chapter 3, effect from correlation processing was maximum when the reference signal is a precise copy of the adopted useful signal. Therefore in channels with the variable parameters for an increase in noise-resistance of reference signal on go side it is desirable to distort so, as was distorted in channel the transmitted signal. Specifically, for this purpose in the examined earlier mutually correlated systems was realized the reconstruction of reference signals in accordance with the results of

the measurement of the parameters of the communication channels. For autocorrelation systems the need for this reconstruction is eliminated, since the role of reference signal plays in essence most adopted useful signal. Therefore autocorrelation systems can sufficiently effectively to work both in the channels with constants and with the variable parameters, including in channels with multiple-pronged propagation. Therefore there is no need for to provide for for receptor the generators of the transmitted signals (or the matched with the transmitted signals filters) and the fairly complicated measuring systems of the parameters of the communication channel.

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Simultaneously with this significantly descend requirements for the accuracy of synchronization.

Thus, the receptors of the autocorrelation broadband systems can be realized considerably simpler than mutually correlated ones. Furthermore, since in the receiver of autocorrelation system there is no reconstruction of the taken signals, as the transmitted signals in

such systems can be used not only the pseudorandom broadband signals, obtained from the special fairly complicated generators, as this occurred in mutually correlated systems, but also the cuts of noise from the "natural" sources: thyratrons, noise diodes, etc. The application/use of such generators for shaping the transmitted in the assigned band signal frequencies also substantially simplifies the construction of the transmitters of autocorrelation systems. At the same time the presence in the reference signal of the autocorrelation system not only of the useful taken signal, but also the noises of the communication channel increases interference level at the output/yield of correlator. Furthermore as it will be shown below, such systems possess sufficiently high inherent noise level (the so-called system interferences). All this lowers the potential interference rejection of autocorrelation systems. The energy lossses of autocorrelation systems in comparison with mutually correlated comprise with the probabilities of errors $p \geq 10^{-4}$ not less than 9-15 dB.

Thus, comparative simplicity of the realization of autocorrelation systems is reached by the value of a decrease in their potential interference rejection. Therefore the practical realization of such systems can be justified when the considerations of simplicity of the construction of receptor are prevailing.

Furthermore, the application/use of autocorrelation systems can turn out to be advisable and then, when the damage of the structure of the transmitted signals in the communication channel are so/such great, thatthe reconstruction of received signals and, consequently, also work of systems of the type mutually correlated cannot be any satisfactory.

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§4.2. Fundamental principles of the construction of autocorrelation systems.

The autocorrelation reception of signals provides for processing in the receptor of the adopted sum of useful signal and interferences of the communication channel with the aid of the diagram of autocorrelator. The block diagram of this correlator is represented in Fig. to 4.2.1 and includes the following in principle necessary cell/elements: multiplier, equipment/device (line) of delay and integrator (see §2.3). The input of correlator enters the sum of useful signal and interferences $x(t)$, which is introduced on multiplier directly, also, through the delay line for a period τ. The

output voltage of multiplier $x(t) \cdot x(t-\tau)$ is integrated for time T. As a result the output voltage of diagram with the sufficiently large interval of integration turns out to be proportional to the autocorrelation function of the received signal:

$$R(\tau)=\lim_{T\to\infty}\frac{1}{T}\int_0^T x(t)\,x(t-\tau)\,dt. \qquad (4.2.1)$$

However, the emission of useful signal out of this voltage without the special construction of the transmitted signal largely it is impeded due to the interference effect of the communication channel. Let us explain this as follows. Let the input of correlator enter the useful noise-like signal $\tilde{z}(t)$ with the uniform in sufficiently broadband F spectrum $S_{\tilde{z}}(f)$, and also fluctuating interference $\xi(t)$ out of the channel of the communication/connection, which has the zero average value and the uniform in the band of frequencies F spectrum $S_{\xi}(f)$.

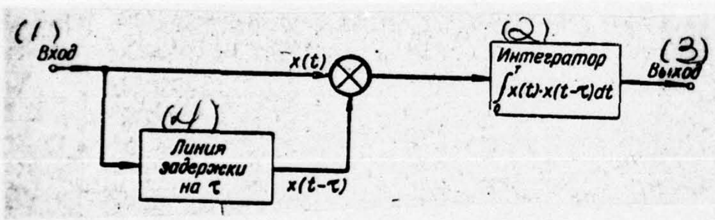

Fig. 4.2.1.

Key: (1). Input. (2). Integrator. (3). Output/yield. (4). Delay line on.

The spectra of signal and interference are shown in Fig. 4.2.2, in which f_{cp} is the medium frequency of the spectra, which

considerably exceeds their value ($f_{cp} \gg F$). Received signal x (t) is equal of this case to the sum of the obtained signal $\tilde{z}$ (t) and interference ξ (t), but its output potential of correlator according to (4.2.1) composes value

$$R(\tau)=\lim_{T\to\infty}\frac{1}{T}\int_0^T\left[\tilde{z}(t)+\xi(t)\right]\left[\tilde{z}(t-\tau)+\xi(t-\tau)\right]dt=$$
$$=\lim_{T\to\infty}\left[\frac{1}{T}\int_0^T\tilde{z}(t)\,\tilde{z}(t-\tau)\,dt+\frac{1}{T}\int_0^T\tilde{z}(t)\,\xi(t-\tau)\,dt+\right.$$
$$\left.+\frac{1}{T}\int_0^T\xi(t)\,\tilde{z}(t-\tau)\,dt+\frac{1}{T}\int_0^T\xi(t)\,\xi(t-\tau)\,dt\right]=$$
$$=R_{\tilde{z}}(\tau)+R_{\tilde{z}\xi}(\tau)+R_{\xi\tilde{z}}(\tau)+R_{\xi}(\tau), \quad (4.2.2)$$

where $R_{\tilde{z}}(\tau)$ and $R_{\xi}(\tau)$ are autocorrelation functions of respectively useful signal and interference; $R_{\xi\tilde{z}}(\tau)$ and $R_{\tilde{z}\xi}(\tau)$

are their mutually correlated functions.

Let us assume that signal $\tilde{z}(t)$ and interference $\xi(t)$ are stationary and statistically independent, since are caused by completely different sources. Furthermore, the average value of interference $\overline{\xi(t)}$ = with 0. Then from mutually correlated functions are equal to

$$R_{\tilde{z}\xi}(\tau)=R_{\xi\tilde{z}}(\tau)=0. \qquad (4.2.3)$$

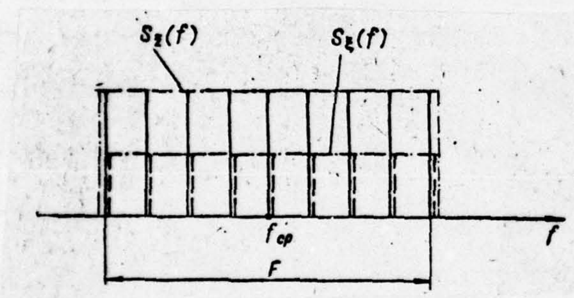

Fig. 4.2.2. Page 195.

In this case from formula (4.2.2) it follows that the autocorrelation function of received signal is equal to the sum of the

autocorrelation functions of useful signal and interference:

$$R(\tau)=R_{\tilde{z}}(\tau)+R_{\xi}(\tau). \qquad (4.2.4)$$

Autocorrelation function $R_{\tilde{z}}(\tau)$ limited in the band of frequencies F of signal with the uniform spectrum, as this was shown in §2.5, takes the form of the harmonic oscillations of frequency f_{cp} with envelope, that is changed according to the law $\sigma^2_{\tilde{z}} \frac{\sin\frac{F\tau}{2}}{\frac{F\tau}{2}}$. Here $\sigma^2_{\tilde{z}}$ is dispersion (power) of useful signal $\tilde{z}$ (t). Plotted function $R_{\tilde{z}}(\tau)$ depending on τ is represented in Fig. 4.2.3a. As can be seen from the figure, the main maximum of envelope (range of powerful correlation) is arranged between values $\pm 1/F$. Remaining maximums, attenuating proportionally $F\tau$, rapidly they decrease in value. At values $\tau > \frac{2\div 3}{F}$, the autocorrelation function of signal $\tilde{z}$ (t) is virtually equal to zero. The autocorrelation function of interference $R_{\xi}(\tau)$ has analogous character with the only difference, that the dispersion (power) of interference σ^2_{ξ} is different from the power of signal $\sigma^2_{\tilde{z}}$.

A change envelope autocorrelation functions $R_{\tilde{z}}(\tau)$ and $R_{\xi}(\tau)$ is shown by dotted line in Fig. 4.2.3b. In this same figure solid

line represented dependence on τ the correlation function of received signal R (τ), which for each fix/recorded τ is equal to the sum of the corresponding values of the autocorrelation functions of useful signal and interference. From the figure one can see that the functions $R_{\tilde{z}}(\tau)$ and $R_{\xi}(\tau)$ are combined in the case in question on the axis τ. For different τ the values of the autocorrelation function of interference $R_{\xi}(\tau)$ are proportional to the appropriate values of the autocorrelation function of useful signal $R_{\tilde{z}}(\tau)$. The more the component of useful signal in the output voltage of correlator, the more the component of interference, and vice versa. At $\tau = 0$ is reached the maximum of the stress of useful signal and output voltage of correlator; however, in this case the disturbing voltage also has the greatest value.

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Thus, as a result of the fact that in the case in question autocorrelation functions $R_{\tilde{z}}(\tau)$ and $R_{\xi}(\tau)$ are combined on the axis τ, the interferences of the communication channel substantially affect the output voltage of correlator, largely impeding the reception of useful signal.

For the elimination of the indicated phenomenon in broadband systems is substantially necessary the superimposition on the axis τ in the method of the autocorrelation functions of useful signal and interference. The indicated superimposition can be reached by the introduction of artificial delays into the transmitted signals. For this explanation let us turn to Fig. 4.2.4 on which is represented block/module/unit the diagram of formation (coding of the transmitted signals of autocorrelation system.

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The transmitter of Fig. 4.2.4 contains as the coding equipment/device the generator of the broadband noise-like signal z (t), delay line for a period τ_1 and summator. The spectrum of signal z (t) is uniform and limited in the band of frequencies F, but its autocorrelation function is similar to the correlation function of Fig. 4.2.3a. The delay time in the line τ_1 is selected large in comparison with time of the correlation of signal 1/F. Virtually it suffices to select it equal to 2-3/F.

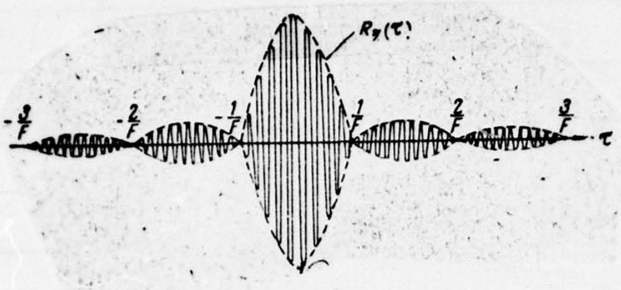
The transmitted signal z_1 (t) is form/shaped at the output/yield of the summator of the coding equipment/device and contains component z (t) and z $(t-\tau_1)$. The first their base is remove/taken directly from the output/yield of generator, and the second from the output/yield of delay line. The autocorrelation function of the transmitted signal $R_{z1}(\tau)$ takes the form

$$R_{z1}(\tau)=\lim_{T\to\infty}\frac{1}{T}\int_0^T z_1(t)\,z_1(t-\tau)\,dt. \quad (4.2.5)$$

Whence, taking into account that $z_1(t) = z(t) + z(t-\tau_1)$, we will obtain

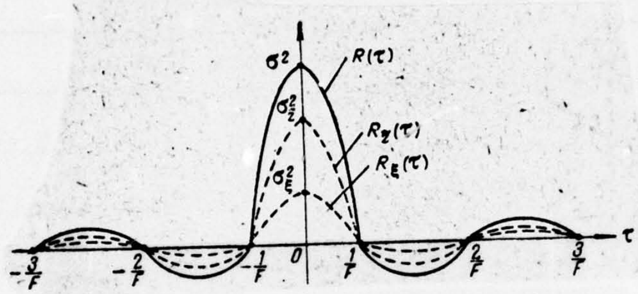
$$R_{z1}(\tau)=\lim_{T\to\infty}\Big[\frac{1}{T}\int_0^T z(t)\,z(t-\tau)\,dt+\frac{1}{T}\int_0^T z(t-\tau_1)\times \times z(t-\tau)\,dt+\frac{1}{T}\int_0^T z(t)\,z(t-\tau_1-\tau)\,dt+ +\frac{1}{T}\int_0^T z(t-\tau_1)\,z(t-\tau_1-\tau)\,dt\Big]=R_z(\tau)+ +R_z(\tau-\tau_1)+R_z(\tau_1+\tau)+R_z(\tau-\tau_1+\tau_1)= =2R_z(\tau)+R_z(\tau-\tau_1)+R_z(\tau_1+\tau), \tag{4.2.6}$$

where $R_z(\tau)$ - the autocorrelation function of signal $z(t)$.



$R_y(\tau)$
$-\frac{3}{F}$
$-\frac{2}{F}$
$-\frac{1}{F}$
$\frac{1}{F}$
$\frac{2}{F}$
$\frac{3}{F}$
τ

a)



σ^2
$R(\tau)$
σ_z^2
$R_z(\tau)$
σ_ξ^2
$R_\xi(\tau)$
τ
$-\frac{3}{F}$
$-\frac{2}{F}$
$-\frac{1}{F}$
0
$\frac{1}{F}$
$\frac{2}{F}$
$\frac{3}{F}$

b)

Fig. 4.2.3.

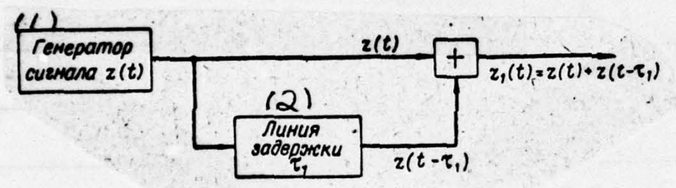


Fig. 4.2.4.

Key: (1). Signal generators. (2). Delay line.

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From equality (4.2.6) it is evident that the autocorrelation function of the transmitted signal is determined by three terms, each of which depends on the form of the autocorrelation function of signal $R_z(\tau)$,

and also of the values of the delay time τ_1 and of the parameter τ.

The character of change $R_{z1}(\tau)$ from τ at the selected value $\tau_1 > \frac{2 \div 3}{F}$ is shown in Fig. 4.2.5. at values $\tau \leqslant \frac{1}{F}$ the main role in the formation of the autocorrelation function of the transmitted signal plays the term $2R_z(\tau)$, that has maximum with τ = 0. The value of maximum is determined by the dispersion (with a power) of entire transmitted signal σ_{z1}^2. The contribution of other two terms is negligible. When changes τ within limits $\tau_1 - \frac{1}{F} \leqslant \tau \leqslant \tau_1 + \frac{1}{F}$ the fundamental contribution introduces second term $R_z(\tau - \tau_1)$, and $2R_z(\tau)$ and $R_z(\tau+\tau_1)$ are negligible. Maximum $R_z(\tau - \tau_1)$ occurs with $\tau = \tau_1$. It is formed because of component z $(t-\tau_1)$ and has a value, equal to the half of the dispersion of entire transmitted signal. As concerns the third term, $R_z(\tau+\tau_1)$ very little at any values τ.

Thus, when the transmitted signal is form/shaped of two components of z (t) and of z $(t-\tau_1)$, its current-regulating function has two maximums, arrange/located one relative to another at a distance selected delay time τ_1.

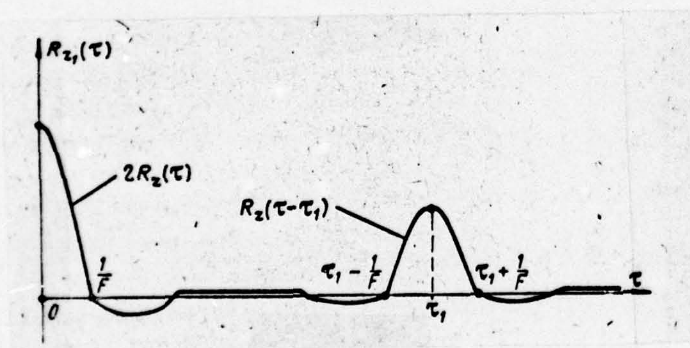
$R_{z_1}(\tau)$
$2R_z(\tau)$
$R_z(\tau-\tau_1)$
$\frac{1}{F}$
$\tau_1 - \frac{1}{F}$
$\tau_1 + \frac{1}{F}$
τ
0
τ_1

Fig. 4.2.5.

Because of this automatically is realized the "superimposition" in the receptor of the autocorrelation functions of useful signal and fluctuating interferences of the communication channel. Actually, during the transmission of signal z_1 (t) = z (t) + z $(t-\tau_1)$ to the input of decoder (autocorrelator) Fig. 4.2.1 will enter the undergone distortion in the communication channel useful signal $\tilde{z}_1$ (t) = $\tilde{z}$ (t) + $\tilde{z}$ $(t-\tau_1)$.

Discussing analogous with the derivation of expression (4.2.6), it is not difficult to show that the autocorrelation function of this signal takes the form

$$R_{\tilde{z}1}(\tau)=2R_{\tilde{z}}(\tau)+R_{\tilde{z}}(\tau-\tau_1)+R_{\tilde{z}}(\tau+\tau_1), \qquad (4.2.7)$$

where $R_{\tilde{z}}(\tau)$ is a short-term autocorrelation function of each component of signal $\tilde{z}_1$ (t). At the same time, the components of $\tilde{z}$ (t) and of $\tilde{z}$ $(t-\tau_1)$ of useful signal are proportional to the appropriate components of the transmitted signal. Then autocorrelation function for $\tilde{z}_1$ (t), shown in Fig. 4.2.6a, repeats on the determined scale the correlation function of the transmitted signal. It also has two maximums, arrange/located with time interval τ_1. The first of them occurs at values τ from 0 to 1/F and is form/shaped because of term $2R_{\tilde{z}}(\tau)$

from (4.2.7), i.e., because of both components of $\tilde{z}(t)$ and of $\tilde{z}(t-\tau_1)$ of the taken signal. The second maximum exists for $\tau_1 - \frac{1}{F} < \tau < \tau_1 + \frac{1}{F}$ and it is determined by term $R_{\tilde{z}}(\tau-\tau_1)$ from (4.2.7), i.e., only by one component of $\tilde{z}(t-\tau_1)$ of entire useful signal. Let us note that if the distortions in channel were absent $(\tilde{z}_1(t)=z_1(t))$, then the autocorrelation function of the taken useful signal corresponded precisely the curve of Fig. 4.2.5.

Further, the autocorrelation function of the interferences of the channel of communication/connection $R_\xi(\tau)$, which have that which was limited in the band of frequencies F and the uniform spectrum, is arrange/located, as this was shown above, in essence in range of values τ from 0 to 1/F and with $\tau > \frac{2 \div 3}{F}$ it is very insignificant (Fig. 4.2.6b).

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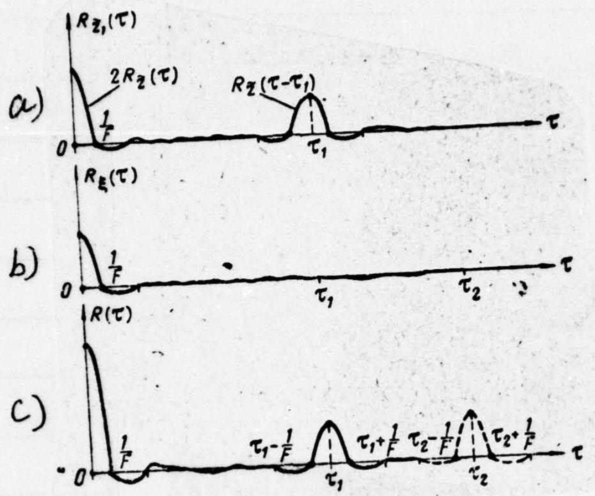
The autocorrelation function of entire received signal, determined by relationship/ratio (4.2.4), is equal to the sum of the autocorrelation functions of useful signal $R_{\tilde{z}_1}(\tau)$ and of interference $R_\xi(\tau)$ (Fig. 4.2.6c). It is not difficult to see that

the interference ξ (t) exerts a substantial influence only on the first of its maximums. The second of them, arrange/located in range of values τ from τ_1 - 1/F to τ_1 + 1/F, is largely released from the effect of interference ξ (t).

Of this consists the proposed in work [21] principle of the superimpositon of the correlation functions of useful signal and interferences of the decoder of autocorrelation communicating system. A difference in the in practice utilized in autocorrelation systems systems of coding and decoding lies in the fact that the formation transmitted and processing received signals in them is realized not during infinite time, but in the finite interval of the duration of the cell/element of the signal T, whose value is determined by the speed of transmission of discrete report/communication. Then the output potential of correlator depends on used realization of signal in the finite interval of duration T and is determined already by the short-term autocorrelation function of the signal

$$X(\tau)=\frac{1}{T}\int_0^T x(t)\,x(t-\tau)\,dt. \qquad (4.2.8)$$



$R_{\xi_1}(\tau)$
$2R_\xi(\tau)$
$R_\xi(\tau-\tau_1)$
a)
$\frac{1}{F}$
0
τ_1
τ
$R_\xi(\tau)$
b)
$\frac{1}{F}$
0
τ_1
τ_2
τ
$R(\tau)$
c)
$\frac{1}{F}$
0
$\tau_1-\frac{1}{F}$
$\tau_1+\frac{1}{F}$
$\tau_2-\frac{1}{F}$
$\tau_2+\frac{1}{F}$
τ_1
τ_2
τ

Fig. 4.2.6.

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In this respect, value X (τ) is unlike (4.2.1) the random variable, which is characterized by certain by average on all possible realizations of signal value $\overline{X(\tau)}$ and by dispersion $\overline{[X(\tau) - \overline{X(\tau)}]^2}$. However, if the transmitted signals possess ergodic property, and the same are many signals of radio communication, then $\overline{X(\tau)}$ it is the autocorrelation function R (τ) from (4.2.1), and the root-mean-square deviation $\sqrt{\overline{[X(\tau)-\overline{X(\tau)}]^2}}$ characterizes the measure of the scatter of the short-term autocorrelation function of its relatively average value.

In this case the expressed above positions remain valid, characterizing the operating principle of autocorrelation systems on the average independent of the selected on each cell/element realization of signal. The special feature/peculiarities of the freedom from interference of autocorrelation systems, connected with the measurement of short-term correlation function (4.28), are discussed in the following paragraph. Here let us note that the effective separation of the first and second maximums of the short-term autocorrelation function of received signal (4.2.8) is possible only, in the first place, when using signals with

sufficiently broadband F, whose time of correlation many times is shorter than the duration of the cell/element of signal, i.e., with

$$\frac{1}{F} \ll T. \quad (4.2.9)$$

In the second place, delay time τ_1 in transmitter for component $\tilde{z}(t-\tau_1)$ must two or three times exceed time of the correlation:

$$\tau_1 > \frac{2 \div 3}{F}. \quad (4.2.10)$$

It is obvious that the signal of form $z_1(t) = z(t) + z(t-\tau_1)$ can be used as one of the two transmitted signals in binary communicating system, for example as the signal of premise/impulse (pressure).

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For the formation/education of the signal of pause (release) is sufficient to set in equipment/device of coding another delay time τ_2, to which will correspond the transmitted signal $z_2(t) = z(t) + z(t - \tau_2)$. In this case the autocorrelation function of received signal $\tilde{z}_2(t) = \tilde{z}(t) + \tilde{z}(t - \tau_2)$ also will have two maximums, one of which is subjected to the powerful effect of interferences and is arrange/located in range of change τ from 0 to $1/F$, but the second to a considerable degree is excess this and is arrange/located in range $\tau_2 - \frac{1}{F} < \tau < \tau_2 + \frac{1}{F}$. The position of the second maximum is shown in Fig. 4.2.6c by dotted line, whereupon for a certainty τ_2 was taken as longer than τ_1. The separation of maximums during transmission $z_2(t)$ is possible also only during satisfaction of the condition

$$\tau_2 \geqslant \frac{2 \div 3}{F}. \qquad (4.2.11)$$

Furthermore, for the discrimination of the signals of premise/impulse and pause the second maximums with τ_1 and τ_2 must be divided by interval (see Fig. 4.2.6c)

$$|\tau_2 - \tau_1| \geqslant \frac{2 \div 3}{F}. \qquad (4.2.12)$$

Relationship/ratios (4.2.9) - (4.2.12) are the conditions, superimposed on the utilized in autocorrelation broadband system signals.

Recall that the correlation function of received signal, determined by relationship/ratio (4.2.8), is the dependence of the output voltage of decoder from the value of time τ its delay line. It is obvious that for recording received signals $\tilde{z}_1$ (t) or $\tilde{z}_2$ (t) it is necessary to select value τ in receiver, equal to a signal delay in transmitter, i.e., $\tau = \tau_1$ for signal $\tilde{z}_1$ (t) and $\tau = \tau_2$ for $\tilde{z}_2$ (t). Respectively receiver in this case must contain two decoders with delay lines on τ_1 in one of them and on τ_2 in other. the installation of the corresponding delay time in transmitter can be realized with the aid of line with removal/outlets on τ_1 and τ_2, switched in accordance with the work of telegraph. Therefore autocorrelation

broadband systems with modulation (manipulation) of the supplementary maximum of the correlation function of the transmitted signals are called another the systems of korrelyatsionno- time/temporary modulation.

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The principle of the construction of the broadband autocorrelation transmission systems of discrete information presented can be explained with the aid of examples. As the first example let us examine the work of the system, proposed to Lange and Muller [21, 49]. The block diagram of its transmitter is represented in Fig. 4.2.7 and includes the coding equipment/device, and also cascade/stages of the frequency conversion and power gain of the output signals of transmitter. The coding equipment/device consists of the noise generator, of band-pass filter, line of delay and summator. As noise generator is utilized the simplest generator, which can be carried out, for example, on thyratron or noise diode. voltage from the output/yield of noise generator is passed through the band-pass filter with band F and the medium frequency f_{cp}, which considerably exceeds the value of band $(f_{cp} \gg F)$. The fundamental requirement, presented for noise generator and filter, is the

provision uniform in the band of frequencies F of the spectrum of signal z (t) at the output/yield of filter. Furthermore, value 1/F must satisfy condition (4.2.9). The correlation function of this limited on noise bandwidth takes the form of the harmonic oscillations of frequency f_{cp} with envelope, that is changed according to the law sine $\frac{\sin\frac{F\tau}{2}}{\frac{F\tau}{2}}$ (for example, see Fig. 4.2.3a).

Fig. 4.2.7.

Key: (1). Coding equipment/device. (2). Noise generator. (3). Band-pass filter. (4). Frequency converter. (5). power amplifier. (6). Delay line.

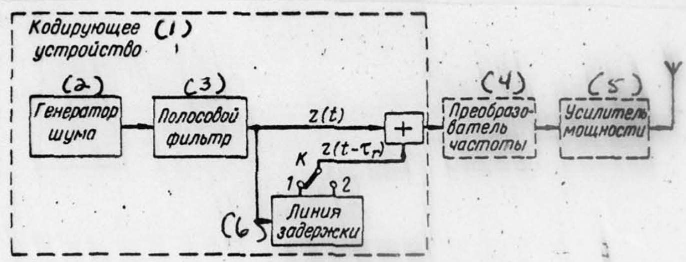

Fig. 4.2.7.

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The fundamental maximum of envelope (range of powerful correlation) is arranged in range $\tau = \pm 1/F$. remaining maximums rapidly decrease in value, so that at values $\tau > \frac{2 \div 3}{F}$ the autocorrelation function of signal is virtually equal to zero. Further from the noise voltage z (t) in the coding device is realized the formation of the which are subject to transmission signals of premise/impulse and pause respectively:

$$\left.\begin{aligned} z_1(t) &= z(t) + z(t-\tau_1); \\ z_2(t) &= z(t) + z(t-\tau_2). \end{aligned}\right\} \qquad (4.2.13)$$

For this z (t) it is supplied to adder directly, also, through the delay line. Line has two removal/outlets, that ensure delay for a period τ_1 for the signal of premise/impulse z_1 (t) and τ_2 for the signal of pause z_2 (t). The commutation of removal/outlets is realized by key/wrench K in accordance with the work of the transmitting telegraph.

Delay factors τ_1 and τ_2 are selected by such that they would satisfy conditions (4.2.10) - (4.2.12). For the duration of premise/impulse from telegraph, equal by T, the duration of the formed signals composes value $T+\tau_1$ and $T + \tau_2$ respectively. In order that upon the exchange of the cell/elements of signal (for example, premise/impulse to pause and vice versa) would not occur a noticeable change in the power level of the emitted signals, value τ_1 and τ_2 besides the indicated above conditions must simultaneously satisfy the also following condition:

$$\tau_1;\ \tau_2 \ll T. \qquad (4.2.14)$$

It is not difficult to see that in this case the system in question is system with active pause. Receptor is its represented in Fig. 4.2.8. It includes the common/general/total input circuits of receiver (high-frequency amplifiers, converters, IF amplifiers), the decoders of the circuits of processing the signals of premise/impulse z_1 (t) and pauses z_2 (t), and also comparison circuit (subtractor).

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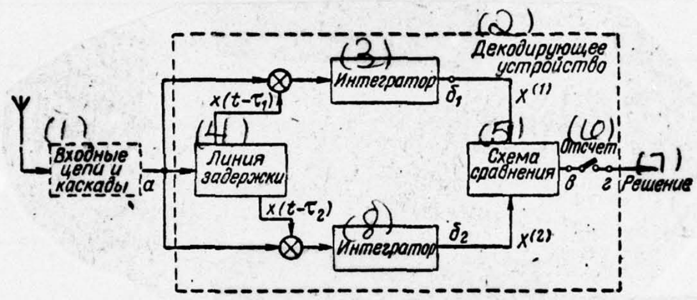
Each decoder is the autocorrelator, which consists of delay line (for a period τ_1 for signal z_1 (t) and τ_2 for signal z_2 (t)), a multiplier and an integrator. The adopted signal.

$$x(t) = \tilde{z}(t) + \tilde{z}(t-\tau_r) + \xi(t),\ r = 1;\ 2, \qquad (4.2.15)$$

Intermediate frequency enters the input of the multipliers of the

circuits of processing received signals directly, also, through the delay line. Let, for example, was transmitted signal z_1 (t). Then the input of the multiplier of the circuit of processing signal z_1 (t) (matched circuit) enter voltage $x(t) = \tilde{z}(t) + \tilde{z}(t - \tau_1) + \xi(t)$ and $x(t - \tau_1) = \tilde{z}(t - \tau_1) + \tilde{z}(t - 2\tau_1) + \xi(t - \tau_1)$. As a result on multiplier prove to be the introduced into synchronism components of $\tilde{z}(t - \tau_1)$ of signals x (t) and $x(t - \tau_1)$, i.e., the component, not delayed in transmitter and delayed in receiver, and the component, delayed in transmitter, but not delayed in receiver. Because of these components is form/shaped basic part of the voltage of the second maximum of autocorrelation function with $\tau = \tau_1$ on output of circuit (Fig. 4.2.9a). According to (4.2.8) and (4.2.15) this voltage is equal

$$\begin{aligned} X^{(1)} = & \frac{1}{T}\int_0^T \tilde{z}(t-\tau_1)\tilde{z}(t-\tau_1)\,dt + \frac{1}{T}\int_0^T \xi(t)\xi(t-\tau_1)\,dt + \\ & + \frac{1}{T}\int_0^T \tilde{z}(t)\tilde{z}(t-\tau_1)\,dt + \frac{1}{T}\int_0^T \tilde{z}(t)\tilde{z}(t-2\tau_1)\,dt + \\ & + \frac{1}{T}\int_0^T \tilde{z}(t-\tau_1)\tilde{z}(t-2\tau_1)\,dt + \frac{1}{T}\int_0^T \tilde{z}(t)\xi(t-\tau_1)\,dt + \\ & + \frac{1}{T}\int_0^T \xi(t-\tau_1)\tilde{z}(t-\tau_1)\,dt + \frac{1}{T}\int_0^T \xi(t)\tilde{z}(t-\tau_1)\,dt + \\ & + \frac{1}{T}\int_0^T \xi(t)\tilde{z}(t-2\tau_1)\,dt. \end{aligned} \qquad (4.2.16)$$



Декодирующее устройство
Интегратор
$x(t-\tau_1)$
Входные цепи и каскады
Линия задержки
δ_1
$X^{(1)}$
Схема сравнения
Отсчет
Решение
$x(t-\tau_2)$
Интегратор
δ_2
$X^{(2)}$

Fig. 4.2.8.

Key: (1). Input target/purposes and cascade/stages. (2). Decoder. (3). Integrator. (4). Delay line. (5). Comparison circuit. (6). Reading. (7). Solution. (8). Integrator.

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First term in expression (4.2.16) forms basic part of the maximum of voltage during $\tau = \tau_1$. Second term is the result of the effect of fluctuating interference. The third, fourth and fifth terms are determined by the fact that the transmitted signal consists of two (delayed and not delayed) components. These terms form the so-called system interference. And finally, the latter of four terms are caused by the interaction of interference and components of signal. All terms, except the first, form the inherent noise of autocorrelation system. the voltage $X^{(2)}$ in the second (unmatched) circuit of processing with $\tau = \tau_1$ and to transmission z_1 (t) is also random variable and is determined by relationship/ratic (see Fig. 4.2.9b)

$$X^{(2)} = \frac{1}{T}\int_0^T \left[\tilde{z}(t) + \tilde{z}(t-\tau_1) + \xi(t)\right]\left[\tilde{z}(t-\tau_2) + \right.$$
$$\left. + \tilde{z}(t-\tau_1-\tau_2) + \xi(t-\tau_2)\right] dt. \quad (4.2.17)$$

In this voltage with the selected values τ_1 and τ_2 are absent synchronous components. As can be seen from (4.2.17) and in this circuit is an increased inherent noise level of system. Specifically, as a result of the presence of these noises autocorrelation systems prove to be more badly on the potential interference rejection than mutually correlated.

Further during the transmission of signal $z_2(t)=z(t)+z(t-\tau_2)$ processing received signal $x(t)=\bar{z}(t)+\bar{z}(t-\tau_2)+\xi(t)$ is realized analogously, but in this case occurs the maximum of

voltage $X^{(2)}$ with $\tau = \tau_2$ (see Fig. 4.2.9b - dotted line) because of the synchronism of the components of form $\tilde{z}(t - \tau_2)$. The stress level in that which was matched and mismatched circuits is determined in this case by expressions (4.2.16) and (4.2.17) with the appropriate replacement in them of index "1" by "2" and vice versa.

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Voltages $X^{(1)}$ and $X^{(2)}$ from the output/yield of decoders enter the comparison circuit, as which can be used subtractor. At the torque/moment of the termination of the cell/element of signal (torque/moment of reading) with

$$X^{(1)} - X^{(2)} > 0 \tag{4.2.18}$$

is accepted the solution to transmission $z_1(t)$. In the case

$$X^{(1)} - X^{(2)} < 0 \tag{4.2.19}$$

is recorded signal z_2 (t).

Figure 4.2.10 shows dependence in time of the voltages at the different points of diagram in Fig. 4.2.8 during the transmission of

the sequence of signals z_1, z_2, z_1.

Another example of system with korrel4qionno-time/temporary modulation is the system, at whose transmitter is analogous with diagram in Fig. 4.2.7, and decoders of receiver, presented in Fig. 4.2.11, contains the only one correlator with delay time τ_1.

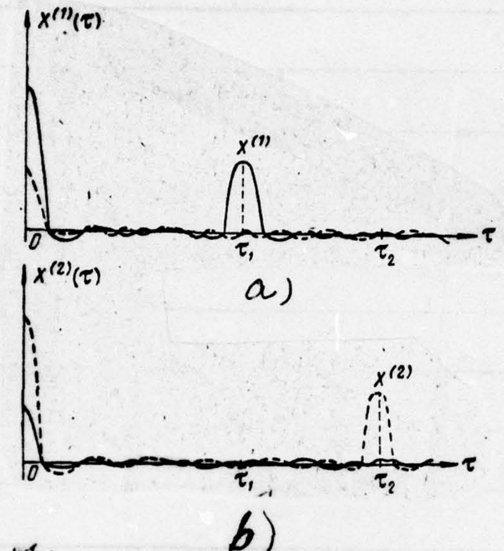

Fig. 4.2.9.

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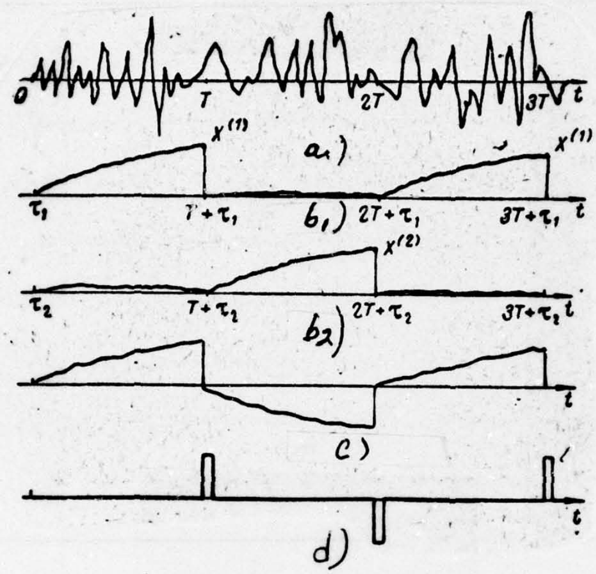
The voltage, which corresponds to the supplementary maximum of correlation function, is isolated in this receiver only during the transmission of signal z_1 (t). Another transmitted signal z_2 (t) at the output/yield of correlator is absent and creates only noise voltage. The output voltage of correlator at the torque/moment of the termination of the cell/element of signal enters the threshold device with level A. If output voltage exceeds value A, then is accepted the solution to transmission z_1 (t), otherwise is recorded z_2 (t).



0
T
2T
3T
t
$x^{(1)}$
a)
$x^{(1)}$
τ_1
$T+\tau_1$
b_1)
$2T+\tau_1$
$3T+\tau_1$
t
$x^{(2)}$
τ_2
$T+\tau_2$
$2T+\tau_2$
$3T+\tau_2$
t
b_2)
t
c)
t
d)

Fig. 4.2.10.

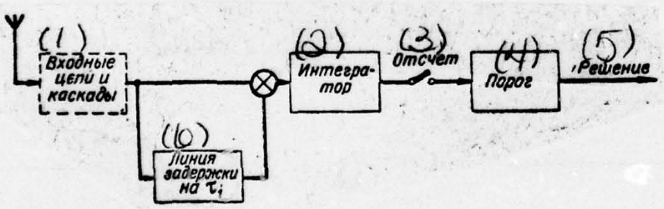


Fig. 4.2.11.

Key: (1). Input circuits and cascade/stages. (2). Integrator. (3). Reading. (4). Threshold. (5). Solution. (6). Delay line on.

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The value of threshold A is different in this diagram from zero and is selected in order to ensure the minimum probability of error. In

this case they say that in receiver is established/installed the optimum threshold level.

It is not difficult to see that this system is system with passive pause. Its advantage is certain simplification in the receiver circuit and the decrease in the inherent noise level of system. At the same time in this case usefully is utilized the only half of the energy taken by a transmitter of signal. This fact, and also the presence of that noted from zero threshold levels lead to the fact that the freedom from interference of this system proves to be somewhat below (approximately to 3 dB) in comparison with the examined in the preceding/previous example system with active pause.

As an example of system with the reduced inherent noise level, but the effective use of an energy of entire transmitted signal can serve the presented in Fig. 4.2.12 autocorrelation system with opposite signals [9]. Its distinctive special feature/peculiarity is the fact that the coding device of transmitter includes the only one delay line of the fixed/recorded time τ, and also phase inverter of 180°.

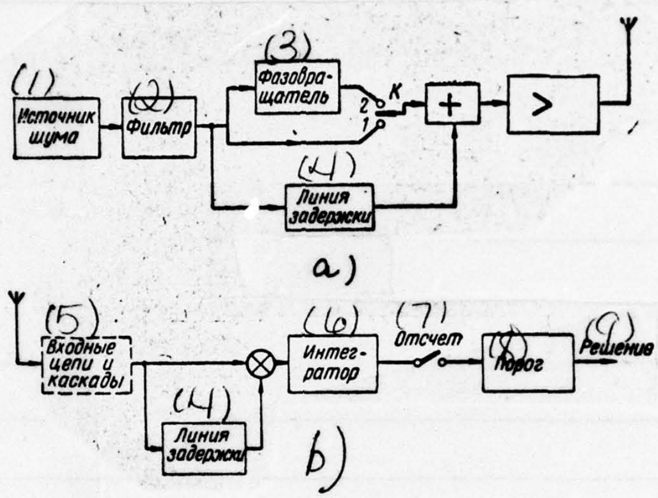
(1)
Источник шума
(2)
Фильтр
(3)
Фазовращатель
К
2
1
+
>
(4)
Линия задержки
a)
(5)
Входные цепи и каскады
(4)
Линия задержки
(6)
Интегратор
(7)
Отсчет
(8)
Порог
(9)
Решение
b)

Fig. 4.2.12.

Key: (1). Noise source. (2). Filter. (3). Phase inverter. (4). Delay line. (5). Input circuits and cascade/stages. (6). Integrator. (7). Reading. (8). Threshold. (9). Solution.

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The undelayed component of the form/shaped signals enters the summator either through the phase inverter or directly in accordance with the regulation of key/wrench K, switched by the premise/impulses of telegraph. The transmitted signals take the form:

$$\left.\begin{aligned} z_1(t) &= z(t) + z(t-\tau); \\ z_2(t) &= -z(t) + z(t-\tau). \end{aligned}\right\} \qquad (4.2.20)$$

The decoder of receiver switches on one correlator, useful output of potential of which it is form/shaped because of component $\frac{1}{T}\int_0^T \tilde{z}(t-\tau)\tilde{z}(t-\tau)\,dt$ during transmission $z_1(t)$ and component - $\frac{1}{T}\int_0^T \tilde{z}(t-\tau)\tilde{z}(t-\tau)\,dt$ during transmission $z_2(t)$. Consequently, these voltages

at the
torque/moment of the reading have different polarities. They enter the threshold device, the optimum value of threshold level of which in this case equal to zero. If the output voltage of correlator is positive, then is recorded signal z_1 (t), otherwise - signal z_2 (t). This system possesses the highest freedom from interference of all examined diagrams.

§4.3. On the potential interference rejection of autocorrelation systems with korrel4qionno-time/temporary modulation.

We will consider the potential interference rejection of the diverse variants of systems with korrel4qionno-time/temporary modulation. Let us assume that the noise generator of transmitter in all these diagrams generates the normal noise z (t) with the uniform spectrum in the band of frequencies F a to the input of receptor except useful signal it enters additive interference ξ (t) in the form cf normal white noise in the band of frequencies of useful signal. We consider also that by means of the selection of the corresponding value F time of correlation is much shorter than the duration of cell/element T of the transmitted signals:

$$\frac{1}{F} \ll T. \tag{4.3.1}$$

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Furthermore, the values of delay time are taken so that they satisfy the inequalities:

$$\left.\begin{aligned}\frac{2\div 3}{F}\leqslant\tau_1;\ \tau_2\ll T,\\ |\tau_1-\tau_2|\geqslant\frac{2\div 3}{F}.\end{aligned}\right\}\qquad(4.3.2)$$

Let us turn at first to the decisive diagram in Fig. 4.2.8, being limited to the case, when the transmitted signals z_1 (t) = z (t) + z (t - τ_1) and z_2 (t) = z (t) + z (t - τ_2) are equiprobable.

The rule of the solution in the diagram in question lies in the fact that signal z_1 (t) is recorded, if the output voltages of the circuits of processing $X^{(1)}$ and $X^{(2)}$ satisfy the inequality

$$X^{(1)} - X^{(2)} > 0. \qquad (4.3.3)$$

With opposite inequality sign is recorded signal z_2 (t). Let us assume that transmits signal z_1 (t). Then the probability of error is probability that inequality (4.3.3) is not fulfilled. In this inequality $X^{(1)}$ and $X^{(2)}$ they are the random variables, determined by relationship/ratios (4.2.16) and (4.2.17). utilizing expansions for signal $\tilde{z}$ (t) and interferences ξ (t) in Fourier series in range from 0 to T,

$$\left.\begin{aligned} \tilde{z}(t) &= \sum_{k=k_1}^{k_2} (A_k \cos k\omega_0 t + B_k \sin k\omega_0 t); \\ \xi(t) &= \sum_{k=k_1}^{k_2} (a_k \cos k\omega_0 t + \beta_k \sin k\omega_0 t). \end{aligned}\right\} \qquad (4.3.4)$$

where $A_k, B_k, \alpha_k, \beta_k$ are independent normal random values with the average values $\overline{A}_k = \overline{B}_k = \overline{\alpha}_k = \overline{\beta}_k = 0$ and dispersions $\overline{A}_k^2 = \overline{B}_k^2 = \frac{P_0}{2FT}$; $\overline{\alpha}_k^2 - \overline{\beta}_k^2 = \frac{\nu^2}{T}$ (P_0 is power of the useful received signal; ν^2 - the spectral density of fluctuating interference).

Then during satisfaction of conditions (4.3.1) and (4.3.2) it is possible to show that the voltages $X^{(1)}$ and $X^{(2)}$ are the independent normal random variables, which have at transmission $z_1(t)$ the average values

$$\left.\begin{aligned}\overline{X^{(1)}} &= \frac{\nu}{2T} h^2;\\ \overline{X^{(2)}} &= 0\end{aligned}\right\} \tag{4.3.5}$$

and the dispersions

$$\left.\begin{aligned}\sigma^{(1)^2} &= \frac{\nu^4}{4T^2}\left(\frac{5}{2}\,\frac{h^4}{FT}+4h^2+2FT\right);\\ \sigma^{(2)^2} &= \frac{\nu^4}{2T^2}\left(\frac{h^4}{FT}+2h^2+FT\right).\end{aligned}\right\} \quad (4.3.6)$$

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In these expressions: FT is a base of system $h^2=\frac{P_0T}{\nu^2}$ - the ratio

of the energy of the taken useful signal to the spectral density of fluctuating interference. Consequently, value $X=X^{(1)}-X^{(2)}$ also is normal random variable with the average value and dispersion, respectively:

$$\left.\begin{aligned}\overline{X}&=\overline{X^{(1)}}-\overline{X^{(2)}}=\frac{\nu^2}{2T}h^2;\\ \sigma^2&=\sigma^{(1)^2}+\sigma^{(2)^2}=\frac{\nu^4}{4T^2}\left(\frac{9}{2}\frac{h^4}{FT}+8h^2+4FT\right).\end{aligned}\right\} \tag{4.3.7}$$

Then the probability of error during transmission z_1 (t) is equal to

$$p_1=P\{X<0\}=\int_{-\infty}^{0}\frac{1}{\sigma\sqrt{2\pi}}e^{-\frac{(x-\overline{X})^2}{2\sigma^2}}dx. \tag{4.3.8}$$

By substituting in this expression of value $\overline{X}$ and σ^2 from

(4.3.7) and by executing integration, we will obtain

$$p_1=\frac{1}{2}\left[1-\Phi\left(\frac{h^2}{\sqrt{\frac{9}{2}\frac{h^4}{FT}+8h^2+4FT}}\right)\right], \tag{4.3.9}$$

where $\Phi(x)$ - probability integral (function of Crump), tabulated, for example, in [3, 35].

It is analogous, during the transmission of signal $z_2(t)$, $\overline{X}^{(1)}$ and $\overline{X}^{(2)}$ is normal and independent variables random variable with the following parameters:

$$\left.\begin{aligned}&\overline{X^{(1)}}=0;\ \overline{X^{(2)}}=\frac{\nu^2}{2T}h^2;\\ &\sigma^{(1)2}=\frac{\nu^4}{2T^2}\left(\frac{h^4}{FT}+2h^2+FT\right);\ \sigma^{(2)2}=\\ &=\frac{\nu^4}{4T^2}\left(\frac{5}{2}\frac{h^4}{FT}+4h^2+2FT\right).\end{aligned}\right\} \tag{4.3.10}$$

Then value $X = X^{(1)} - X^{(2)}$ is normal and has an average value and a dispersion respectively

$$X = -\frac{\nu^2}{2T}h^2;\ \sigma^2 = \frac{\nu^4}{4T^2}\left(\frac{9}{2}\frac{h^4}{FT} + 8h^2 + 4FT\right). \tag{4.3.11}$$

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The probability of error during the transmission of signal $z_2(t)$ is equal to the probability of the fulfillment of inequality (4.3.3) and, therefore,

$$p_2 = P\{X > 0\} = \int_0^{\infty} \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{(x-\overline{X})^2}{2\sigma^2}} dx. \quad (4.3.12)$$

Substituting in this expression of value $\overline{X}$ and σ^2 from (4.3.11), it is not difficult to ascertain that probability p_2 takes the analogous (4.3.9) form.

The composite probability of the error of piece-by-piece reception with a priori equal probabilities of the transmitted signals takes the form

$$p = \frac{1}{2}(p_1 + p_2).$$

Then for the system of Fig. 4.2.8 finally we have the following expression for the probability of the error:

$$p = \frac{1}{2}\left[1 - \Phi\left(\frac{h^2\sqrt{FT}}{\sqrt{9h^4 + 16h^2FT + 8(FT)^2}}\right)\right]. \qquad (4.3.13)$$

Let us note that for the case of coherent reception in the optimum mutually correlated system with orthogonal signals the expression for the probability of error took the form

$$p = \frac{1}{2}[1 - \Phi(h)]. \qquad (4.3.14)$$

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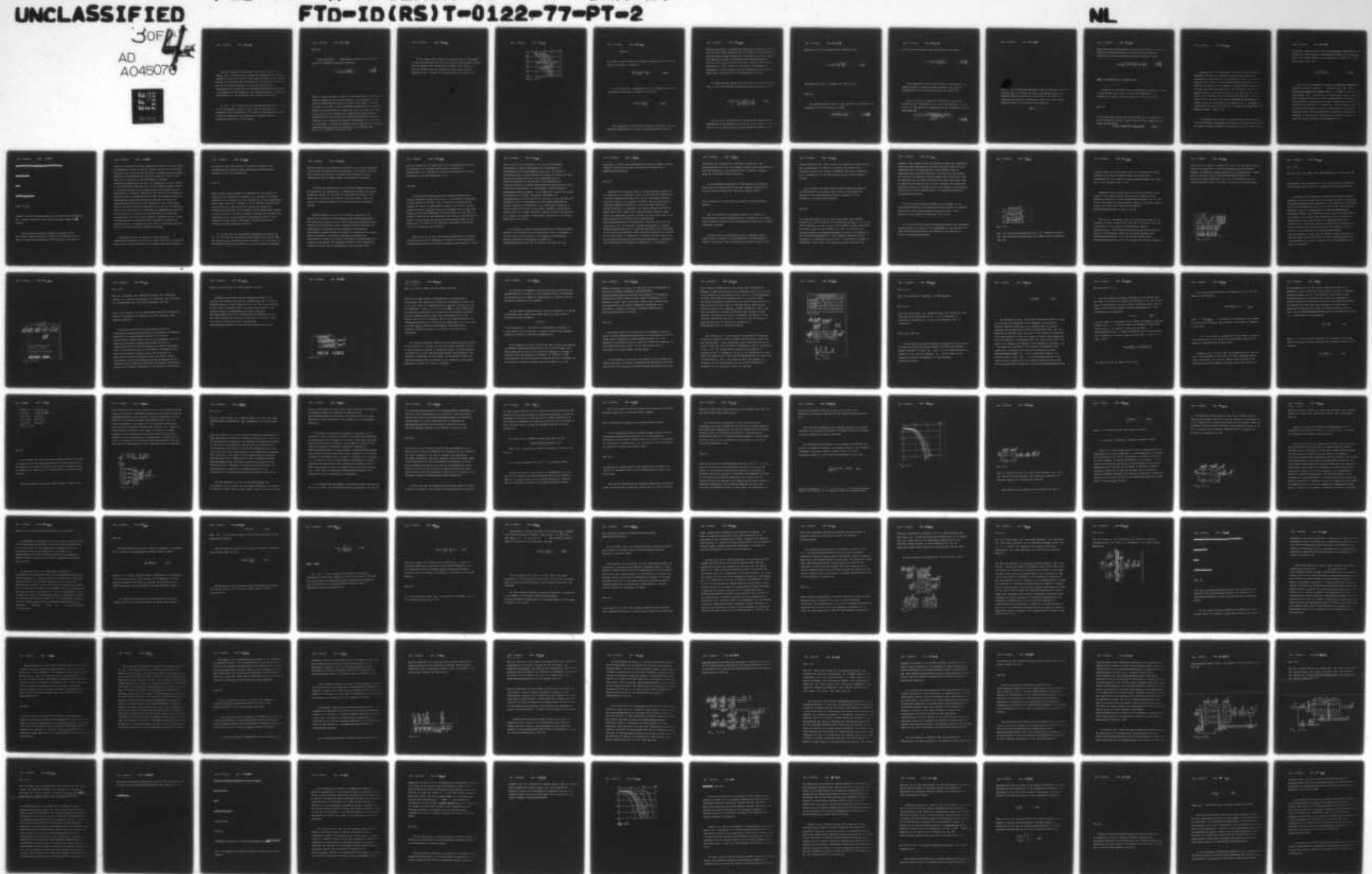

From the comparison of formulas (4.3.13) and (4.3.14) it is evident that in autocorrelation system the probability of the error depends not only on the ratio of the energy of signal to the spectral density of the fluctuating interference h^2, but also on the value of base FT the utilized signals. In this develops itself the nonoptimality of systems with korrel4qionno-time/temporary modulation as consequence of the presence in the reference signal of fluctuating interference ξ (t) and sufficient high inherent noise level.

In Fig. 4.3.1 solid lines are constructed dependences of p on h^2 for the different values of base (10^2, 10^3 and 10^4). The dependence of the probability of the error of autocorrelation system on value FT determines two substantially important special feature/peculiarities of such systems.

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First, with $\frac{h^2}{FT} \gg 1$ and constant value FT from (4.3.13) we cbtain the following asymptotic expression:

$$p \sim \frac{1}{2}\left[1 - \Phi\left(\frac{\sqrt{2FT}}{3}\right)\right], \tag{4.3.15}$$

which it does not depend on value h^2. This means that at sufficiently large in comparison with base FT values h^2 the probability of the error in system depends only on the base of the system: a further increase in the power of the transmitted signals does not improve the freedom from interference of system (see Fig. 4.3.1 with FT = 10^2). At the same time for obtaining the sufficiently small asymptotic probabilities of the errors, which would be considerably than less required in the communicating system of probabilities p = 10^{-4}-10^{-5}, it is necessary to ensure sufficiently large values FT. The virtually indicated requirement will be satisfied, as is evident from Fig. 4.3.1, already at values FT > 10^2.

In the second place, there is an optimum value of base $(FT)_0$, at which the probability of error (4.3.13) in autocorrelation system will be smallest. The optimum value of base can be found, if we (4.3.13) differentiate for parameter FT and, after equating derivative zero, to solve the obtained equation relative to FT.

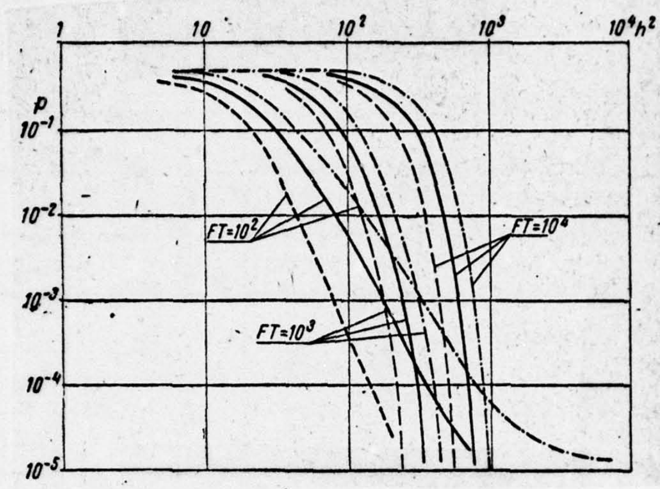

Fig. 4.3.1.

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As a result we will obtain the following expression for the optimum base in the system in question:

$$(FT)_0 = \frac{3}{2\sqrt{2}} h^2 \approx 1,06h^2. \quad (4.3.16)$$

At this value (FT) $_0$ expression (4.3.13), determining potential interference rejection, is converted to the form

$$p \approx \frac{1}{2}\left[1 - \Phi\left(\frac{h}{4}\right)\right]. \quad (4.3.17)$$

From comparison (4.3.17) with (4.3.14) it is evident that with identical probabilities of errors the autocorrelation system in

question plays back by energetically mutually correlated 16 times or on 12 dB. One should emphasize that the freedom from interference of autocorrelation system, determined by expression (4.3.17), requires the regulation of base in accordance with relationship/ratio (4.3.16). This can be realized either by a change in the band of frequencies F or by a change in the speed of transmission of information T. With constant base the probability of the errors in system will be determined by relationship/ratio (4.3.13).

In autocorrelation system with active pause and opposite signals (Fig. 4.2.12) the composite probability of error is equal to [9]

$$p=\frac{1}{2}\left[1-\Phi\left(\frac{h^2}{\sqrt{\frac{5}{2}\frac{h^4}{FT}+4h^2+2FT}}\right)\right]. \tag{4.3.18}$$

In Fig. 4.3.1 by dotted line on (4.3.18) are constructed the dependences of the probability of error on h^2 for cases $FT = 10^2$, 10^3 and 10^4. At the fixed value of base FT and with increase h^2, the

probability of error asymptotically approaches value

$$p \sim \frac{1}{2}\left[1-\Phi\left(\sqrt{\frac{2FT}{5}}\right)\right]. \quad (4.3.19)$$

Probability (4.3.19) is somewhat less than (4.3.15).

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The optimum value of base, at which provides the minimum of the probability of error (4.3.19), is equal

$$(FT)_0 = \frac{\sqrt{5}}{2} h^2 \approx 1,12h^2. \quad (4.3.20)$$

In this case the freedom from interference of the system

$$p \approx \frac{1}{2}\left[1-\Phi\left(\frac{h}{2\sqrt{2}}\right)\right]. \qquad (4.3.21)$$

From comparison (4.3.21) with (4.3.17) it follows that the application/use of opposite signals provides with identical probabilities of the errors energy gain two times (3 dB).

Let us turn now to system with the passive pause whose transmitter is represented in Fig. 4.2.7, and receiver - in Fig. 4.2.11. In this system the expression for the probability of error takes form [29]

$$p=\frac{1}{2}\left\{1+\frac{1}{2}\Phi\left[\frac{2a\sqrt{h^4+2h^2FT+(FT)^2}-h^2\sqrt{2FT}}{\sqrt{5h^4+8h^2FT+4(FT)^2}}\right]-\right.$$
$$\left.-\frac{1}{2}\Phi(a)\right\}, \qquad (4.3.22)$$

where α - the standardized/normalized value of threshold voltage. Let us assume that in receptor the level of threshold voltage α is establish/installed optimum for each state of the channel of communication/connection (depending on value h^2). This optimum threshold value can be found from the condition

$$\frac{\partial p}{\partial \alpha}=0.$$

After fulfilling differentiation (4.3.22) with respect to α and after solving the obtained equation, we will obtain the following expression for the standardized/normalized threshold voltage:

$$\alpha_0 = \frac{1}{h^2}\left[\sqrt{c\left(2FT + \ln\frac{c}{4d}\right)} - 2\sqrt{2dFT}\right], \qquad (4.3.23)$$

where $c = 5h^4 + 8h^2FT + 4(FT)^2$; $d = h^4 + 2h^2FT + (FT)^2$.

In Fig. 4.3.1 dot-dash line are constructed according to formula (4.3.22) taking into account (4.3.23) the dependence of the probability of error p on h^2 for cases FT = 10^2, 10^3 and 10^4.

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At the fixed value of base and with increase h^2, the probability of error, determined (4.3.22) together with (4.3.23), asymptotically approaches the value

$$p \sim \frac{1}{2}\left[1 - \Phi\left(0{,}59\sqrt{FT}\right) + \Phi\left(0{,}338\sqrt{FT}\right)\right]. \qquad (4.3.24)$$

Unfortunately, it is impossible to find the locked analytical expression for $(FT)_0$ in system with passive pause during the simultaneous optimization of its probability of error on FT and α. Numerical value $(FT)_0$ for each concrete/specific/actual h^2 can be found by the methods of the solution to transcendental equations. At the same time about the character of the behavior of system with passive pause it is possible to judge by the given in Fig. 4.3.1 curves of its freedom from interference (dot-dash line). From the figure one can see that this system plays back in the range of the working values of the probability of error $p \geqslant 10^{-4}$ to system with active pause not less than 3 dB, but to system with active pause and opposite signals - about 6 dB.

In conclusion of paragraph it should be noted that the systems with korrel4qionno-time/temporary modulation are weakly shielded from the effect of single harmonic interferences. If in the input voltage

of receptor, besides useful signal and fluctuating interferences, are present the single harmonic interferences of form $z_{\pi}(t)=A_{\pi}\cos(\omega_{\pi}t+\varphi_{\pi})$, that in its output voltage in the composition of value $X^{(r)}$ will enter terms of the type

$$R_{\pi}(\tau_r)=\frac{A_{\pi}^2}{2}\cos\omega_{\pi}\tau_r. \tag{4.3.25}$$

Time of the correlation of harmonic interference considerably exceeds the value of shift τ_r. Therefore term $R_{\pi}(\tau_r)$ has sufficiently large value (nearly equal to the power of the affecting interference $\frac{A_{\pi}^2}{2}$) and "damps" the supplementary useful maximum of the correlation function of signal. Consequently, the presence of single harmonic interference can cause a considerable decrease in the freedom from interference of system with time-correlation modulation. The latter fact impedes the work of such systems in the loaded frequency bands. For the suppression of harmonic interferences it is expedient to utilize their rejection on the input of receptor or to apply the special diagrams of compensation [11].

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Chapter 5. Discrete-address systems of wide-band radio communication.

§5.1. General information about discrete-address communication systems.

Discrete-address radiolink systems are related to the multichannel systems which use a range of VHF. They are of for a military communication/connection large interest, since make it

possible to organize the radio communication between the large number of subscribers, placed on certain, usually limited, the territories during the only condition that the power of transmitters and receiver sensitivity are sufficient for provision within the limits of the datum of the territory of each subscriber's confident communication/connection with each. in widespread at present systems of the multichannel communication, in which common broadband channel is taken also usually in the range of VHF and is condensed by the large number of channels, intended for the joint of two terminal and sometimes several intermediate point/items, the broadband systems of kodovo-address communication/connection provide the conduct of communication/connection with any subscriber of this grid/network with any by other virtually independent of the work of the remaining pairs of the subscribers. the conditions of communication/connection and its quality in this case are obtained such as in usual wire communication in the case of use automatic telephone station. During the application/use of discrete-address systems is provided the effective use of a frequency band and, which is especially important, there is no need for central commutation station.

Characteristic for the systems of discrete-address communication/connection is the use all radio stations of one and the same, usually sufficiently wide, the frequency spectrum, selected in

the range of VHF or DMV without the frequency retuning of the transmitting and receptors, which considerably simplifies both equipment itself and its operation.

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Unique working tuning elements in subscriber are the selector of address and the regulator of loudness, i.e., the same as of usual telephone. As an example let us let us point out to the communicating system RADAS (see §5.4), intended to use in tactical component/link and described into 1964 [43], which provides the conduct of communication/connection in grid/network of 3400 subscribers, placed in territory by 48 X 125 km. In system is utilized the frequency band 300-400 MHz and is allow/assumed the simultaneous conduct of 680 communication/connections. Subscriber's selection and call is realized by simple pushbutton switching.

It is known that the simultaneous transmission of signals of one but that more out of different point/items in one and the same the frequency band difficulties does not cause. At the same time the simultaneous reception of signals in one and the same the frequency

band is connected with essential difficulties due to large dofference in the levels of signals and interferences. On the strength of this the greatest difficulties in designing and realization of such communicating systems appear during the creation of receptors.

As already mentioned in §1.4, the discrete address systems can be synchronous and asynchronous. In synchronous systems are utilized orthogonal signals; in this case the separation of signals in the point/item of reception is realized relatively simply, since the energy of orthogonal signals in principle they can be completely divided.

Typical examples of the use of orthogonal signals are the systems with frequency or time/temporary synchronous packing/seal, the widely being applied in wire, radio relay and tropospheric lines communication/connections. During frequency division multiplex the orthogonality is provided for all possible of the temporary displacements between signals; the spectra of signals do not overlap. During synchronous time-division multiplex is provided the time/temporary diversity of signals, whereupon is allow/assumed the overlap of the spectra. The transient noises are not fundamentally inherent in these forms of packing/seal. Their presence under

practical conditions is caused only by the insufficient accuracy of the realization of equipment (insufficient linearity of cascade/stages, the imperfection of the characteristics of filters, inaccuracy in the synchronization, etc.).

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In asynchronous address systems are applied nonorthogonal (quasi-orthogonal) signals. The signals of the different subscribers in such systems can overlap, also, on time and in frequency. Are possible two cases of nonorthogonal signals. In the first case the signals virtually completely overlap in frequency and on time. Total energy of all mixing subscribers (signals) affects in this case the receiver, which drives one subscriber's reception as interference. An example of this system was already mentioned system RADAS. In the second case the signals of the different subscribers I overlap on time and in frequency only partially.

Central and at the same time complex problem during the design of asynchronous address communicating systems is the separation of nonorthogonal signals on the basis of address sign/criterion. In this

case out of the very principle of the use of nonorthogonal (overlapping) signals escape/ensues the inevitability of interferences or as them frequently they call, the "noises of nonorthogonality", serious difficulties during the creation of such systems they appear also due to the possibility of the formation/education of the signals of all subscribers and formation/education of excess momentum/impulse/pulses because of the conditions of propagation or interferences. Large advantage of asynchronous address systems - in the absence of the need for precise synchronization of system. This makes them simpler on equipment/device and more flexible in operation it is facilitated the admission of the new subscribers into the common communication channel; appears the possibility of the operational control of an entire system with the aid of the special address of interruption; possibly the use of simple repeaters; easily are solved the questions of a change in the number of subscribers, organization of conference service and some others.

The interesting special feature/peculiarity of discrete-address systems is their ability to self-regulation, which is comprised in the fact that with an increase in the number of simultaneously working subscribers deteriorates the quality of communication/connection. But this leads to the fact that the

subscriber, at whose report/communication does not represent special urgency, itself will wait more favorable torque/moment for a communication/connection.

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Emphasizing the positive sides of discrete-address systems, at the same time it is necessary to keep in mind that the provision for a reliable communication/connection with all surrounding stations within the limits of the assigned territory with any point/item of arrangement/permutation proves to be all the same sufficiently difficult problem: in subscriber, who accepts station with the low signal level (moved away), can render/show many neighbors, which create the considerably greater strength of field. This leads to the high probability of obtaining false informational momentum/impulse/pulses (interferences), since the latter will be created even with the imprecise incidence/impingement of foreign momentum/impulse/pulses in the temporary positions, which correspond to the adopted correspondent. Under the actual conditions of the work of the large number of subscribers in limited territory turn out to be necessary supplementary organizational measures for a decrease in such interferences. Specifically, apparently, it is expedient to

divide/mark off territory into individual sections with the removal/diversion of their its frequency ranges, using extensively in this case repeaters. It is expedient also to realize a flexible manner by the powers of radio stations.

In the subsequent paragraphs of this chapter the questions, which relate to realization and the most important properties of discrete-address systems, are examined in greater detail.

§5.2. Formation and the decoding of signals in discrete-address systems.

With the formation of broadband signal in the systems of discrete-address communication/connection, intended for the conduct radio telephone exchange, must be consecutively solved three, largely independent problems.

The first task is transformation of the telephone (vocal) signal, which is subject to the transmission from continuous to discrete. The second task is introduction into this signal of the

address sign/criterion, which ensures the reception of signal only by that correspondent, to whom it is intended. The third task is formation of noise-like signal at subcarrier and carrier frequency, i.e., introduction into the formed pulse group of high-frequency filling.

Let us examine, by which paths are solved these problems. For the discreteness of continuous voice signal is form/shaped the sequence of the momentum/impulse/pulses, following with the determined, so-called clock frequency.

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It is selected usually two or three times higher than greatest frequency out of the spectrum of the modulating voltage, i.e., on the same foundations, as this is made in all systems of pulse modulation. The typical value of this frequency is 8 kHz. In this case the interval between momentum/impulse/pulses is 124 μs. The sequence of the momentum/impulse/pulses of clock frequency undergoes further modulation in accordance with the voltage of the which is subject to transmission voice signal. The greatest practical interest they

represent three possible forms of modulation: phase pulse modulation (pulse position modulation), pulse-code modulation (KIM) and delta modulation (DM). In comparison with other possible forms of modulation pulse position modulation, of KIM and DM, they provide with reception considerably higher freedom from interference, since the amplitude and the shape modulated pulses when using these forms of modulation information will not bear. Recording the transmitted signals in receptor is realized only on the presence or the absence, as this will be shown somewhat below, the appropriate modulated momentum/impulse/pulses.

In the case pulse position modulation with respect to the instantaneous values of the modulating voltage changes the temporary situation of clock momentum/impulse/pulses relative to their positions in the absence of modulation (Fig. 5.2.1).

In the case of KIM to each instantaneous value of the modulating voltage (which it is taken at the torque/moment of the generation of clock momentum/impulse/pulse) corresponds its code sequence of shorter momentum/impulse/pulses.

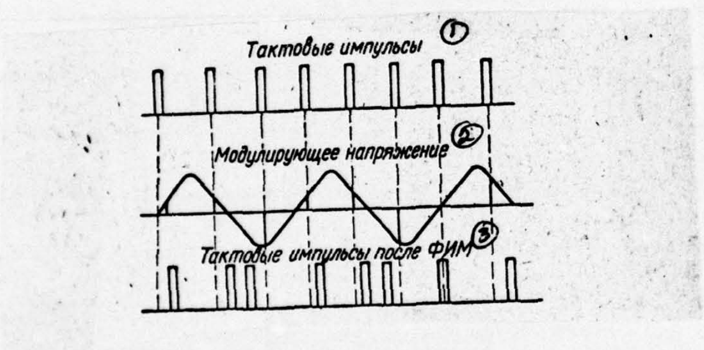

Fig. 5.2.1.

Key: (1). Clock momentum/impulse/pulses. (2). Modulating voltage. (2). Clock momentum/impulse/pulses after pulse position modulation.

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In other words, the instantaneous value of the modulating voltage transmits in the form of binary number, the binary units corresponding to the presence of momentum/impulse/pulse, and binary zero - to its absence (Fig. 5.2.2).

Fundamental pulse repetition rate with KIM is defined as the product of the number of bits, accepted for a coding, for the repetition frequency of clock momentum/impulse/pulses. For the case of five discharges and clock frequency 8 kHz it is obtained equal to 40 kHz; the respectively time interval between momentum/impulse/pulses is equal to 25 µs.

With in. the information about the instantaneous value of the modulating voltage transmits also with the aid of binary code. For a realization in. into system is introduced the special equipment/device, which reveal/detects value and the direction of a change in the modulating stress on the section between clock momentum/impulse/pulses. If on this section between clock momentum/impulse/pulses. If on this section the voltage (envelope) of

signal does not change or changes less than by the determined value, then at the output/yield of delt6ta-modulator goes the periodic sequence of subpulses: regular sequencing of premise/impulse - pause. If on the section between clock momentum/impulse/pulses the modulating voltage grow/rises, then goes the continuous cycle of premise/impulses, while if it drops, there are no premise/impulses.

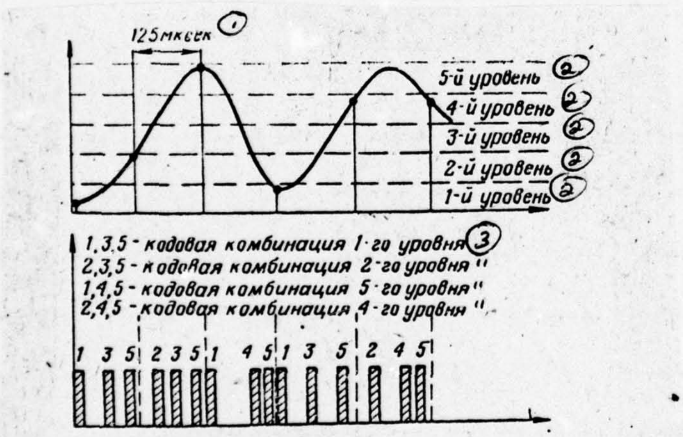

Fig. 5.2.2.

Fig. 5.2.2.

Key: (1). μss. (2). Level. (3). Code combination to level. Page 224.

Schematically this clarified Fig. 5.2.3; there gives the simplified block diagram, which ensures the realization of delta modulation.

A characteristic difference of KIM and DM from pulse position modulation is the fact that with of the first two forms of modulation the momentum/impulse/pulses appear or do not appear at the completely defined torque/moments of time, h that time as with pulse position modulation time of the appearance of a momentum/impulse/pulse previously unknown. This difference can be used on receiving end for an increase in the freedom from interference of the reception of signals KIM and DM, since appears the possibility during the provision for a synchronous working of receiver and transmitter of black-out effect into the pauses between the intervals, when is waited for the appearance of a momentum/impulse/pulse (gating/strobing); at this time the signals of other stations for receiver affect will not be. With the pulse position modulation of this possibility no, since to predict time of the appearance of a pulse is impossible.

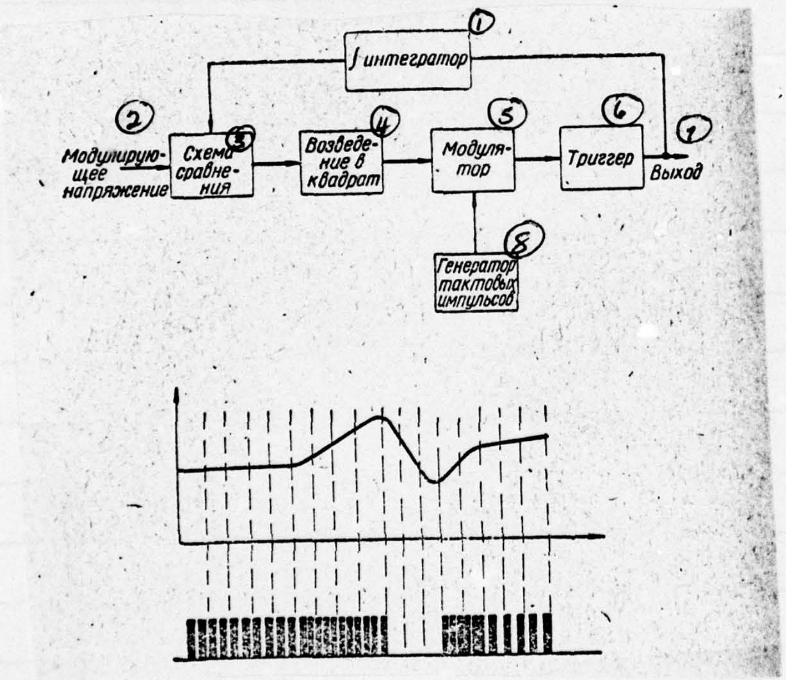
∫ интегратор
Модулирующее напряжение
Схема сравнения
Возведение в квадрат
Модулятор
Триггер
Выход
Генератор тактовых импульсов
1
2
3
4
5
6
7
8

Fig. 5.2.3.

Fig. 5.2.3.

Key: (1). integrator. (2). Modulating voltage. (3). Comparison circuit. (4). Erection into square. (5). Modulator. (6). Flip-flop. (7). Output/yield. (8). Clock pulse generator. Page 225.

Let us note, however, that the synchronous working of receiver with respect to transmitter is necessary alsc for a reception pulse position modulation.

The following task of the signal-shaping circuit of kodovo-address communication/connection is imparting the modulated sequence of the momentum/impulse/pulses of the address sign/criterion, characteristic only for this subscriber. For this purpose each separate momentum/impulse/pulse is coded by transformation into the specific group of new pulses (subpulses), which in the further stages of the formation of broadband signal obtain high-frequency filling; during the introduction of high-frequency filling noise-like signal also obtains address sign/criteria, since high-frequency filling of subpulses for all subscribers can be identical, different for different subscribers, different for different subpulses, it can smoothly or discretely

change on the specific for this subscriber law, etc.

Obtaining the specific group of subpulses (coding) can be realized, for example, with the aid of delay line, with a series of removal/outlets, as this is shown in Fig. 5.24. The general duration of each group of subpulses must be not more than the half of the interval between the neighboring, the initial modulated momentum/impulse/pulses (for a pulse position modulation) and the more whole interval for the case of KIM and DM, of course, taking into account the fact that the intervals among momentum/impulse/pulses in these cases are substantially less.

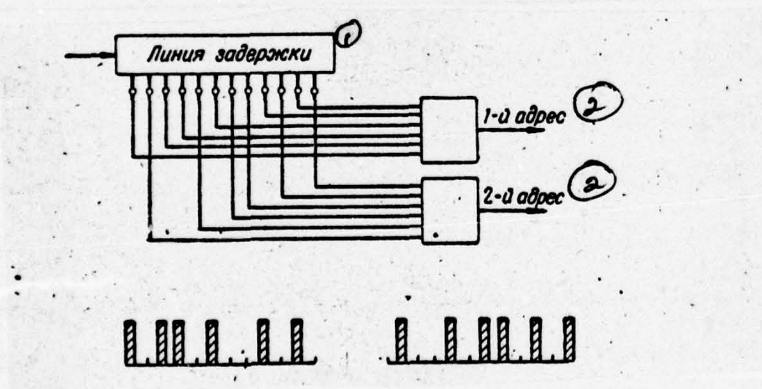

Fig. 5.2.4.

Fig. 5.2.4.

Key: (1). Pine of delay. (2) and address. Page 226.

Figure 5.2.4 shows groups of subpulses for two subscribers of grid/network. The group of 1st subscriber's subpulses is formed from 1, 3, 4, 6, 9 and 11th removal/outlets of the line of delay, but the 2nd subscriber - from 1, 2, 5, 7, 8 and 10-~~go~~th. In these examples for the different subscribers are utilized the only different (besides the 1st) removal/outlets, which, even if desirably, but in the general case it is not necessary; part of the removal/outlets can be common/general/total for a series of the subscribers. As a rule, code is formed only by internal time/temporary pulse separations; therefore 1-1st removal/outlets is utilized in the codes of all subscribers.

The address sequenceof subpulses can be formed also with the aid of the circuits of registers, as this was already clarified in §1.2. In these cases the sequence of subpulses begins to be form/shaped on the admission of clock momentum/impulse/pulse (from modulator). The duration of subpulses and their number in the interval between two momentum/impulse/pulses is assigned by the frequency of the cadence generator, entering the circuit of register.

The versions of addresses (without taking into account of the possibilities of a variation in the frequency filling of subpulses) are determined by the number of cascade/stages in register and are provided by switching in it feedback.

For the concrete/specific/actual circuit of register is a series of versions, which ensure the maximum length of cycle, which is determined by the expression

$$2^n + 2^{n_i} + 1.$$

In this expression n - the number of step/stages of register, n_i - the number of the step/stage, to which is realized the feedback. Formula is valid for a code with foundation two.

As an example let us let us point out that in one of the systems of broadband communication/connection [19], that in somewhat more detail examined in §5.4, in the circuit of register is utilized 17 flip-flops with two versions of feedback - on 3-rd and to 14-th cascade/stages. The first version is utilized for the transmission of the presence of subpulse ("one"), and by the second, i.e., the

absence of subpulse (" zero"). In this case for the transmission of each value of binary information (modulated clock momentum/impulse/pulse) is utilized the sequence of 63 subpulses. Calculations indicate that with this number of subpulses it is possible to ensure 2080 intermittent combinations. However, since some combinations have between themselves considerable time correlation, a number of virtually advisable combinations it decreases approximately two times.

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The further task of the formation of the broadband signal of kodovo- address communication/connection is an introduction into the formed word of subpulses of high-frequency filling, i.e., the formation of broadband signal at carrier or sub-carrier frequency. As has already been spoken above, in this case usually also is introduced into signal address sign/criterion.

Let us examine, as is solved this question in the systems with coding concerning the theory of frequency-time matrix/die, which was some of the first systems of broadband communication/connection and

which obtained propagation in the realized radio communication systems. In these systems each of the subpulses obtains its, one of m of the possible (in the diagram of Fig. 5.2.6 m = 4) high-frequency fillings, which differ in frequency (f_1, f_2, f_3 or f_4). These sub-carrier frequencies resound along the axis of frequencies up to such distances in order that the overlaps between pulse spectra virtually would not be. Modulator circuits chomprise in such a way that the subcarrier voltages frequencies would proceed to group amplifier only during the arrival of subpulse. The signals of all frequency subchannels, and them it will be m, they are united in group amplifier and are emitted by transmitter on common/general/total to carrier frequency.

The totality of the time intervals of subpulses and frequency fillings is accepted to call frequency-time matrix/die (ChVM). For the examined case - three subscribers, seven removal/outlets of the lines of delay and four sub-carrier frequencies - ChVM is represented in Fig. 5.2.5. The general block diagram of the formation of broadband signal is represented in Fig. 5.2.6. Discreteness of voice signal and modulation of clock momentum/impulse/pulses is carried out in cell/elements 1 and 2. The formed by delay line sequence of subpulses obtains high-frequency filling in the cell/elements, designated in the diagram in letters M. Page 228.

	Положения подимпульсов 1									
	1-е	2-е	3-е	4-е	5-е	6-е	7-е	...	...	...
1-й абонент 2	f_1		f_2		f_3					
2-й абонент 2	f_3			f_2		f_4				
3-й абонент 2		f_4		f_1			f_2			
....										

Fig. 5.2.5.

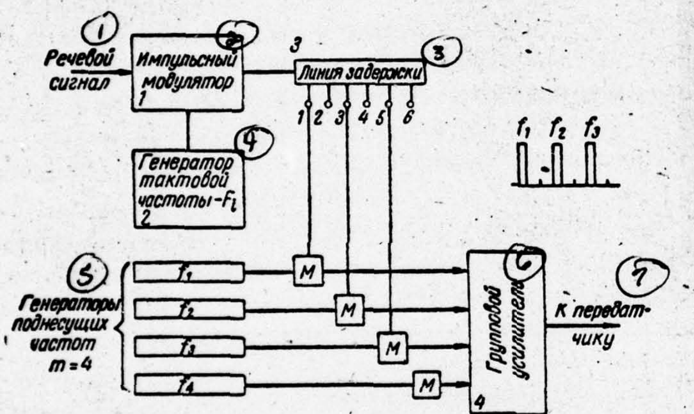

Fig. 5.2.6.

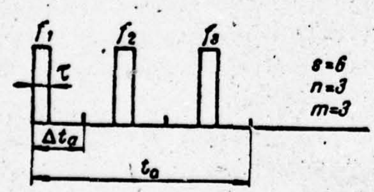

Fig. 5.2.7.

Fig. 5.2.5.

Key: (1). Positions of subpulses. (2). ~~1~~ subscriber.

Fig. 5.2.6.

Key: (1). Voice signal. (2). Pulse modulator. (3). Delay line. (4). Generator clock frequencies are F_i 2. (5). Generators of sub-carrier frequencies m = 4. (6). Group amplifier. (7). To transmitter.

Fig. 5.2.7. Page 229.

Let us examine now some fundamental principles for the circuits of coding in asinxronno-address systems from ChVM. The distance between subpulses in group Δt_a (Fig. 5.2.7) is connected with the duration of the group of subpulses t_a and the number of the temporary positions of subpulses S by the following relationship/ratio:

$$\Delta t_a = \frac{t_a}{S-1}. \tag{5.2.1}$$

The minimum value Δt_a is selected taking into account the need for the clear discrimination of the subpulses, which stand on adjacent temporary positions. If we consider that the passband receiving and transmitting circuits is sufficiently wide, then the expansion of subpulses will occur only in the process of propagation due to the reception of the echo signals and multi-beam propagation. It is not difficult to understand that Δt_a must be more than $\tau + \tau_p$, where τ is a duration of subpulse at the transmitting end/lead, a τ_p - the expansion of subpulse in the process of propagation from transmitter to receiver. The repetition frequencies of clock momentum/impulse/pulses F_i, as has already been spoken, it is selected equal to (2-3) F_{min}, a the displacement/movement of an entire group of subpulses in the process of modulation (with IFM) direct-connected with the period of clock momentum/impulse/pulses and

must not exceed $T_i/2$.

The total number of possible addresses for the special case, when there is the only one sub-carrier frequency, i.e., for the case only of time/temporary coding of address (addresses differ only in terms of time intervals between momentum/impulse/pulses), it is determined by the expression

$$N' = C_{s-1}^{n-1} \qquad (5.2.2)$$

- by the number of combinations of the number of temporary positions without unity (s - 1) from the number of subpulses in group also without one (n - 1), bearing in mind that the first momentum/impulse/pulse is always fix/recorded on the first position. For an example let us let us point out that with s = 6 and n = 3 (Fig. 5.2.8)

$$N' = \frac{(s-1)!}{(n-1)!\,(s-n)!} = \frac{1\cdot 2\cdot 3\cdot 4\cdot 5}{1\cdot 2\cdot 1\cdot 2\cdot 3} = 10.$$

All these versions are shown in Fig. 5.2.8.

If we have several sub-carrier frequencies ($m > 1$), the total amount of the addresses

$$N = C_m^n \sum_{l=0}^{n-1} C_n^l (S-1)^l, \quad (5.2.3)$$

where $C_m^n = \frac{m!}{n!(m-n)!}$ - the number of combinations of the number of sub-carrier frequencies (m) according to the number of subpulses in group (n);

C_n^l - the number of combinaticns of the number of subpulses in grcup according to the number of their possible positions (to number of removal/outlets of delay line).

Formula (5.2.3) is valid under the assumption of the fact that each of the subpulses of this group has their frequency filling (their subcarrier). Let us note, however, that is possible the construction and more complex than ChVM, when on one carrier are

arrange/located not one, but several subpulses of code. It is necessary to keep in mind that from an entire totality of addresses, defined by formula (5.2.2), for practical target/purposes most convenient turn out to be only those, whose all time/temporary interval differ one from another both within the address and between the addresses of this group. Such addresses is conventionally designated as rational. The number of raticnal addresses for the case, when all momentum/impulse/pulses have one frequency filling (m = 1), is determined by the expression

$$N'_{\text{рац}} < \frac{S-1}{C_n^2}. \qquad (5.2.4)$$

When m > 1, by the rational addresses it is accepted to call such, which do not have identical frequency-time intervals. Number of such addresses

$$N_{\text{рац}} < \frac{2S-1}{C_n^2} C_m^2. \qquad (5.2.5)$$

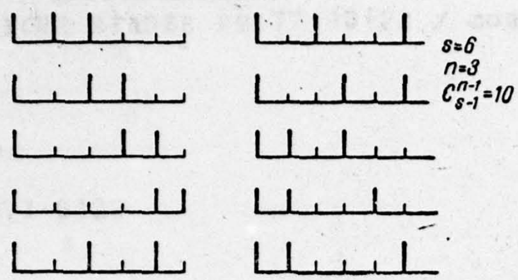

Fig. 5.2.8.

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It is obvious that when using only rational addresses descends the probability of the formation/education of false addresses with the simultaneous arrival of signals from several transmitters they are improved the condition of the discrimination of addresses under receptor.

The block diagram of receiving signal scanner, formed on the

block diagram of Fig. 5.2.6, is given in Fig. 5.2.9. Signals from all working in the given torque/moment subscribers are passed through the common/general/total input amplifier, they are transformed into the signals of intermediate frequency and simultaneously enter m of filters (according to the number of the sub-carrier frequencies), after which are included the delay lines. Each line has a number of removal/outlets according to the number of subpulses in group. Subpulses in the derivations of delay lines after detection are supplied to adder (cascade/stage of agreements), at output/yield of which is form/shaped the resulting momentum/impulse/pulse only when will enter the corresponding to this subscriber series of subpulses.

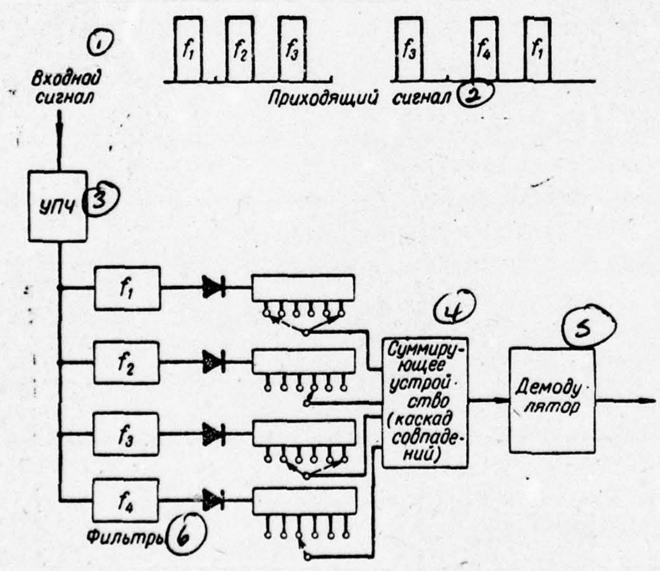

Fig. 5.2.9.

Fig. 5.2.9.

Key: (1). Input signal. (2). Incoming signal. (3). UPCh. (4). Adder (cascade/stage of agreements). (5). Demodulator. (6). Filter. Page 232.

Thus, for instance, if it is necessary to accept the subscriber, to whom corresponds the series of subpulses depicted on Fig. 5.2.6, to the cascade/stage of agreements are supplied the subpulses of channel f_1 from the 6th removal/outlet, channel with f_2 of from the 4th removal/outlet and channel f_3 from the 2nd removal/outlet of delay lines. Upon this connection/inclusion of this subscriber's subpulses they will enter to the inputs of the cascade/stage of agreements simultaneously and at output/yield it will be obtained resultant momentum/impulse/pulse. Upon other connection/inclusions of the removal/outlets of subpulse from delay lines they will enter the cascade/stage of agreements not simultaneously and at its output/yield of momentum/impulse/pulse be obtained it will not be.

For the explanation of that as one should include the cell/elements of the decoder with of another subscriber, the series of subpulse of whom takes the form, shown in Fig. 5.2.9 on top to the

right, in the figure are given dotted lines. After the cascade/stage of agreements usually is included the selector of the momentum/impulse/pulses of duration and further follows the pulse demodulator, which restores the voice signal of the corresponding subscriber.

By examining the work of the schematic of receptor, it is necessary to focus attention on the fact that after the defiltering of common/general/total signal of the input of each delay line enter simultaneously several (according to the number of working in the given torque/moment subscribers) independent sequences of momentum/impulse/pulses. In this totality of momentum/impulse/pulses only a small fraction, namely those, that correspond to the codethe subpulses of subscriber, who is subject to reception, are necessary, i.e., by workers. With the large number of subscribers are always possible such combinations of the nonoperative subpulses which in random combination can form the address code of this radio station, i.e., create false informational momentum/impulse/pulses at the output/yield of the cascade/stage of agreements.

It is natural that the number of such false impulses depends not only on the number of simultaneously working subscribers, but also on

the technical characteristics of the cascade/stage of agreements. The circuit of this cascade/stage can be realized so that those which coincide will consider the momentum/impulse/pulses the coinciding leading edges, or so that coinciding will be considered the momentum/impulse/pulses, which overlap on certain (of course, desirably possibly smaller) to the interval of time and, etc.

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The accuracy of the required agreement will be determined by the accuracy of the circuit parameters, and therefore for the accuracy of the circuit parameters they must be produced are sufficiently stringent requirements. The false impulses, which unavoidably appear at the output/yield of the cascade/stage of agreements, can form false addresses and lead to the noises of nonorthogonality, which relate to so the one who is called to system interferences. These noises usually place the fundamental limitation to an increase in the permissible number of simultaneously working subscribers.

We will show that the formation of noise-like signal to carrier frequency according to the principle of frequency-time matrix/die is

not only possible method. Can find practical application/use also the method, with which are utilized the pseudo-noise signals, shifted in frequency. In this case to each unity of the binary information of this address corresponds the signal by duration T, and the signals of other addresses differ in terms of frequency shift by the value, multiple 1/T.

The signal of the n address in this case takes the form

$$z_n(t) = z(t)\cos[\omega_0 t + n\Delta\omega t + \varphi(t)],\ 0 \leqslant t < T,$$

where ϕ (t) - pseudo-noise signal as duration T, which has as N of bits;

ω_0 - carrier frequency; $\Delta\omega = 2\pi/T$ is a frequency shift.

Pseudo-noise signal ϕ (t) in turn, is form/shaped with delay line or with register. The isolation/liberation of such signals in receptor is realized with the aid of their selection in frequency with the subsequent processing in the discrete matched filters.

Let us note that the indicated pseudo-noise signals did not have extensive application in discrete-address systems.

§5.3. Freedom from interference of discrete-address systems.

In the asynchronous discrete-address systems, intended for the simultaneous communication/connection of the large number of subscribers in the overall frequency band, as a rule, for the purpose of a decrease in charging range, i.e., decrease in the interferences, are applied the systems with passive pause.

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In systems with passive pause by each subscriber's transmitter is emitted the broadband signal of the determined form in alternation with pauses.

The typical representative of broadband system with the passive pause, which uses pulse noise-like signals and the discrete matched

filters, is the system whose description was published into 1965 [19] and data on which were given in §5.4.

The freedom from interference of this system during the transmission of binary information with the aid of premise/impulses and pauses equal prolonged for the case of channel with the constant parameters and the fluctuating interference in transmission of pseudorandom sequence of 63 momentum/impulse/pulses is characterized by curve 1 (Fig. 5.3.1), constructed on the basis of the given in work [19] experimental data.

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Along the Y-axis is here deposit/postponed $h^2 = P_c T/\nu^2$, i.e. the ratio of the energy of the cell/element of signal to the spectral density of the additive fluctuating interference at the input of receptor, while along the axis of ordinates the probability of errors in the reception of discrete information. Let us note that the intelligibility of continuous report/communications (voice signal) is considered satisfactory with incorrect reception not more than two-three cell/elements of 100, in other words, the probability of

errors with reception must not be worse (3-4) • 10^{-2}. This probability of errors is provided for the system in question with $h^2 \geqslant 15$.

Have practical propagation also broadband systems with separate processing the orthogonal components of received signal. The energies of these components are usually identical.

The simplified block diagram of the receptor of system for the case of the quadratic addition of the output voltages of the branches of separate processing signals is given in Fig. 5.3.2. The transmitted signals in this system are represented in the form:

$$\left.\begin{aligned} z_1(t) &= z_1^{(1)}(t) + z_1^{(2)} + \ldots + z_1^{(n)}(t); \\ z_2(t) &= 0, \end{aligned}\right\} \qquad (5.3.1)$$

where the components $z_1^{(i)}$, of i = 1, 2, ..., n of serrated signal satisfy the condition of orthogonality under the intensive sense.

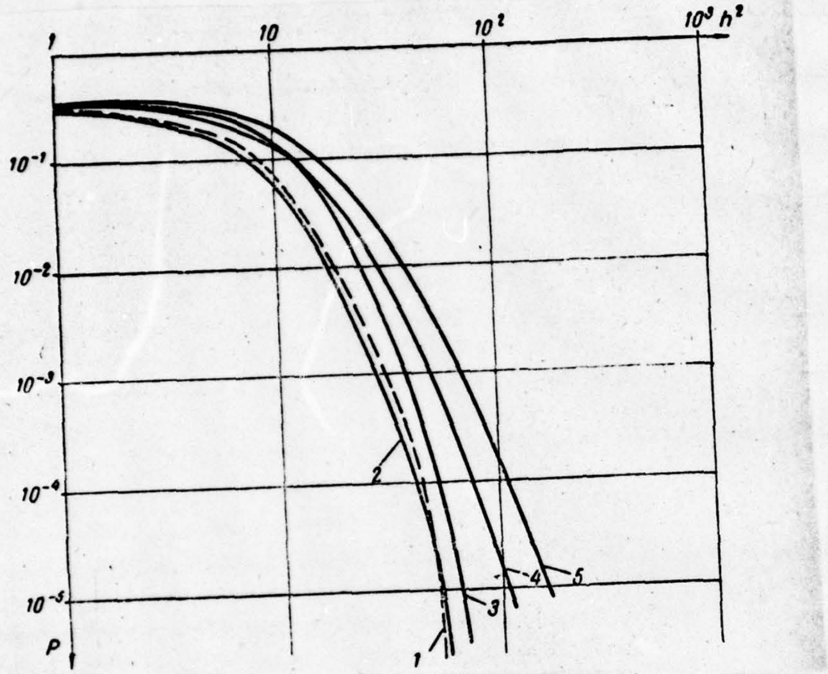

Fig. 5.3.1.

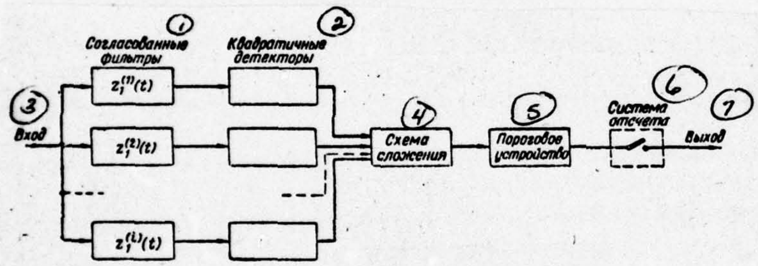

Fig. 5.3.2.

Key: (1). Matched filters. (2). Square law detectors. (3). Input. (4). Circuit of multiplication. (5). Stability thresholds. (6). Reference system. (7). Output/yield. Page 236.

The energies of all components are identical and equal to

$$\frac{1}{n}\int_0^T z_1^2(t)\,dt, \qquad (5.3.2)$$

where T - the duration of the cell/element of signal;

n - the number of branches of separate processing signal.

Figure 5.3.1 gives curves (curves 2 and 3), characterizing the probability of the error depending on h^2 and constructed on the basis of the materials [31]. Curve 2 corresponds to the case, when there are three branches of separate processing (n = 3), curve 3 - to the case of five branches. An increase in the probability of erroneous reception with an increase of the number of branches of separate processing components is caused by the fact that in circuit is applied nonoptimal processing the addition of signals on envelope and because of this occurs the loss of the information about the initial phase of high-frequency filling.

The decisive circuit, given in Fig. 5.3.3., differs from in terms of the examined above fact that the solution to transmission z_1 (t) is accepted only if output potentials of each detector exceed the determined threshold level. Otherwise, if the voltage at least in one of the branches does not exceed threshold level, is accepted the solution to transmission x_2 (t).

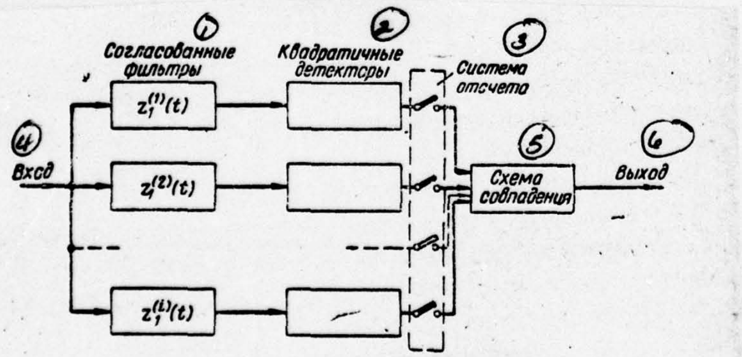

Fig. 5.3.3.

Fig. 5.3.3.

Key: (1). Matched filters. (2). Square law detectors. (3). Reference system. (4). Input. (5). Coincidence circuits. (6). Output/yield.

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Figure 5.3.1 gives the curves of the dependences of the probability of error p on h^2 with $n = 3$ (curve 4) and $n = 5$ (curve 5), constructed also based on materials [31].

As can be seen from the comparison of curves in Fig. 5.3.1, by the best freedom from interference in channel with the constant parameters and the fluctuating interference possess address systems with the discrete matched filters and systems with separate processing components from subsequent quadratic addition. The freedom from interference of such systems is approximately identical. The worst freedom from interference possess systems with the reading of the voltage in each branch of processing. The energy loss of such systems with respect to the first composes at the probability of error 10^{-4} value of approximately 5 dB. From the results of the examination of noise-resistance it follows also that in systems with separate processing is expedient the use of a relatively small number of components ($n \leq 5$), i.e., the signals of a comparatively simple form. An increase in the amount of components can cause a noticeable

decrease in the freedom from interference of such systems.

In conclusion of paragraph, let us note that an essential effect on the freedom from interference of asinxronno-addresssystems have not only the noises of the communication channel, but also the interferences, are, as was noted above, the noises of the nonorthogonality of those utilized by different correspondents signals. These noises limit the number of simultaneously utilized radio stations.

The given in Fig. 5.3.1 curves makes it possible to rate/estimate the permissible number of simultaneously working radio stations depending on the level of their interferences, fluctuating noise and base of the utilized signals. Let M be a number the simultaneously working in this band F radio stations. Let us assume that the power and the durations of signals of all radio stations are identical and equal to respectively P_c and T. Furthermore, let us consider that value M is sufficiently great, in consequence of which the resulting interference, formed (M - 1) by those mixing signals, is normal random process with zero mathematical expectation and dispersion $\sigma^2_{\text{вп}} = \nu^2_{\text{вп}} F$, where $\nu^2_{\text{вп}}$ - the spectral density of interference.

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As shown in work [7], with the indicated assumptions in practice for all utilized in axynchronously addressed systems signals

$$\nu^2_{\text{ВП}} = \frac{\alpha(M-1)P_0}{F}, \qquad (5.3.3)$$

where $\alpha \gg 1$ - certain correction factor, determined by the structure of the utilized signals, more precise, by the properties of their mutually correlated functions; for perfect (in the sense of the smallest possible value of mutually correlated function) signals $\alpha = 1$.

As a result of the presence of interferences with spectral density (5.3.3) the resulting quantity of spectral noise density

$$\nu^2 = \nu_0^2 + \nu_{\text{ВП}}^2, \tag{5.3.4}$$

where ν_0^2 is the spectral density fluctuating interference of the communication channel.

Then the value of the ratio of the energy of signal to spectral noise density takes the form

$$h^2 = \frac{P_c T}{\nu^2} = \frac{P_c T}{\nu_0^2 + \nu_{\text{ВП}}^2}. \tag{5.3.5}$$

By realizing in this expression simple transformations taking into account (5.3.3), let us present finally value h² in the following form:

$$h^2 = \frac{h_0^2}{\left[1 + \frac{a(M-1)h_0^2}{FT}\right]}, \qquad (5.3.6)$$

where $h_0^2 = \frac{P_0 T}{\nu_0^2}$.

Let h^2_{TP} be the required for obtaining the assigned probability of error (for example, 10^{-4} or 10^{-5}) value h^2. Then from (5.3.6) we obtain the permissible number simultaneously working in the assigned band F of asinxronno-address systems, which is determined by the relationship/ratio

$$M_{\text{доп}} \leqslant \frac{1}{\alpha}\left(\frac{FT}{h_{\text{тр}}^2} - \frac{FT}{h_0^2}\right) + 1. \quad (5.3.7)$$

From this formula it is evident that the value $M_{доп}$, first, is directly proportional to the base of the utilized signals FT, i.e., the permissible number of simultaneously working radio stations grow/rises with an increase, for example, in the band of the utilized signals.

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In the second place, value $M_{доп}$ the properties of signals, i.e., at the tendency of factor $1/\alpha$ to one.

Furthermore, as can be seen from (5.3.7), value $M_{доп}$ depends on the relationship/ratio between $h^2_{тр}$ and h^2_0. If $h^2_{тр} = h^2_0$, that $M_{доп} = 1$, but if $h^2_0 \to \infty$, that maximally possible number of utilized in this band radio stations

$$M_{доп\,max} \leqslant \frac{1}{\alpha}\frac{FT}{h^2_{тр}} + 1. \qquad (5.3.8)$$

Let, for example, $FT = 10^3$, $\alpha = 1$ and $h^2_{тр} = 45$, which corresponds to the probability of the error of order 10^{-4} in systems with the discrete matched filters or with quadratic addition. Then $M_{доп\,max} \leqslant 24$.

The more detailed information about the procedure of calculation of the number of permissible simultaneously working asinxronno-address transmissions in the assigned band F can be found in works [7, 40 and 53].

§5.4. Practical systems of broadband discrete-address communication/connection.

At the present time in the foreign periodic literature published large number of materials about the discrete-address systems, developed by different firms.

These systems are very diverse in their designation/purpose, in the methods of formation and decoding of broadband signal, in their technical characteristics they are recommended both for the military and for the civil application/use. In the majority of the cases the detailed characteristics and the parameters of systems on are given and publication have to high degree advertising character. Was not establish/installed the terminology, characterizing or classing already obtained the realization of system.

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To the basin in all after some broadband discrete-address systems were establish/installed the following mostd widely use designations:

RACEp - Random Access Correlation for Extended Performance - the system of arbitrary subscriber's call with correlation for an improvement in the characteristics; RADAS - Random Access Discrete Adress System is a system of arbitrary subscriber's call by discrete address; RADEM - Random Access Delta Modulation - the system of arbitrary subscriber's call with delta modulation.

System, which uses a coding according to the principle of ChVM (RASEP). Materials about this discrete-address radio-telephone communicating system were published in 1961 [46, 47]. The system works in the range of metric and decimeter waves and provides the simultaneous work of 70 subscribers on 35 duplex channels, but taking into account the fact that the transmitters do not work continuously, i.e., the coefficient of the use of channels (set/assuming it equal to 0.1), to 700 subscribers without essential interferences. With a further increase in the number of simultaneously working subscribers is observed an essential increase in the interferences and a communication/connection becomes unsatisfactory. For a decrease in the interferences in system is provided for the disconnection of transmitters for a period of pauses in conversation. For all channels is utilized common/general/total frequency channel 4 MHz wide. To system it was developed in several versions. Movable the version of equipment provides range to 3.2 km, and portable —— to 25 km. The

firm, which developed this system, indicates the large number of possible versions of its use for the civil and military target/purposes.

The simplified block diagram of transmitter is given in Fig. 5.4.1. The momentum/impulse/pulses from clock pulse generator, modulated on position (pulse position modulation), come the course of delay line, from removal/outlets of which after the admission of each clock momentum/impulse/pulse is remove/taken the series of the subpulses, the time intervals between which are determined by the selected removal/outlets from the lines of delay. The numbers of the selected removal/outlets are switched in special equipment/device and are the parameter of address code.

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These subpulses open/disclose the gates, which pass to group voltage amplifier with a frequency of f_1, f_2, f_3 from the separate oscillators. The frequencies f_1, f_2, f_3 can be establish/installed by subscriber and also are one of the parameters of address. At the output of group amplifier are obtained high-frequency subpulses at

different frequencies in accordance with the taken frequency-time matrix/die, i.e., by called subscriber's address code. As an example Fig. 5.4.1 shows ChVM for two subscribers. Respectively in the switching equipment/device is shown commutation for the 1st subscriber (solid lines) and for the 2nd subscriber (dotted line).

The block diagram of receiving part is given in Fig. 5.4.2.

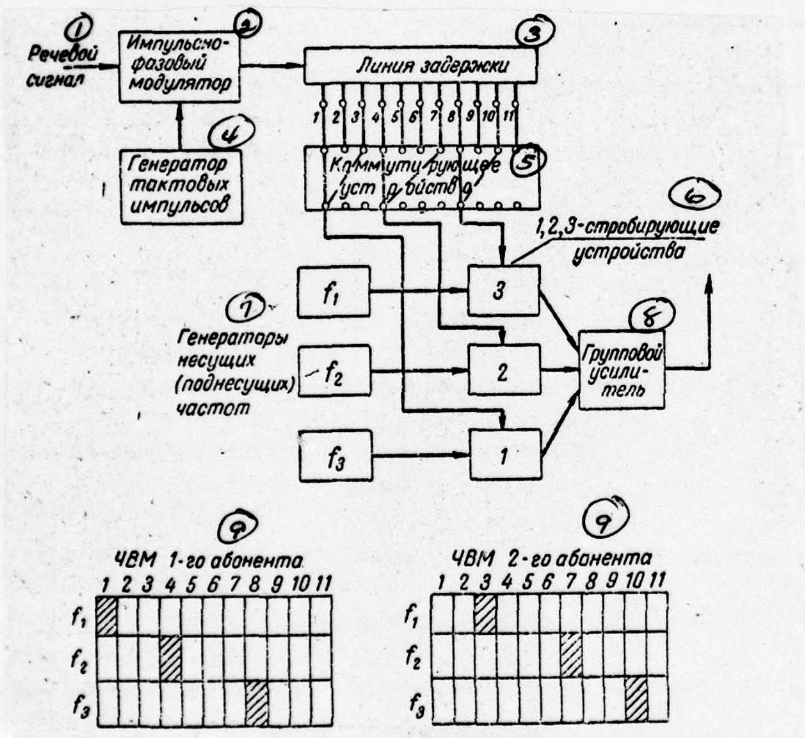

Fig. 5.4.1.

Fig. 5.4.1.

Key: (1). Voice signal. (2). Pulse-phase modulator. (3). Delay line. (4). Clock pulse generator. (5). To switching equipment/device. (6). 1. , 2. , 3. gates. (7). Generators of carriers (sub-carrier frequencies). (8). Group amplifier. (9). ChVM the 1st subscriber.

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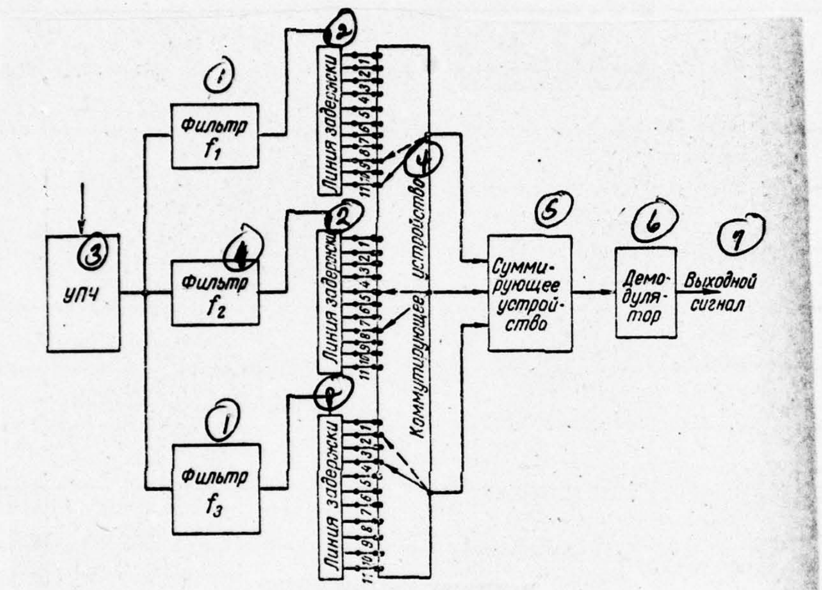
The taken and converted to intermediate signal frequency enter three filters - through the number of frequencies ChVM. After filters and the decoding of subpulses are supplied to the inputs of three lines of delay. Delay lines are included mirror to the inclusion of delay line in the circuit of formation at the transmitting end/lead, as this was clarified in §1.1. Subpulses in the removal/outlets of delay lines through the switching equipment/device are supplied to adding circuit (cascade/stage of agreements), salient output pulse only when not its input simultaneously come momentum/impulse/pulses from all delay lines. Necessary correspondent's reception is provided by the conformity of filter frequencies to oscillator frequencies at the transmitting end/lead (f_1, f_2, f_3) and by the commutation of removal/outlets from delay lines to adder. On the block diagram of Fig. 5.4.2 in the switching equipment/device are shown two cases of commutation for the reception of two subscribers with respect to two cases in Fig. 5.2.1.

Fig. 5.4.2.

Key: (1). Filter f_1. (2). Delay line. (3). UPCh. (4). Switching equipment/device. (5). Adder. (6). Demodulator. (7). Output signal.

end section.



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From adder the momentum/impulse/pulses, which correspond to the modulated clock momentum/impulse/pulses, are supplied to the pulse-phase demodulator, which isolates the initial modulating signal.

In this system for each subscriber are provided for three address codes. If necessary to cause and to conduct the exchange of

information with another subscriber is utilized the address of this subscriber's call. If necessary for correspondent's urgent call, who in the given torque/moment works with other, is provided for the address of interruption, also, if necessary for the establishment of simultaneous communication/connection with all correspondents - the address of circular call.

Synchronous sampled-data system. Below broadband synchronous communicating system [43] is intended for the radio-telephone exchange between movable objects with the distances between them to 8 km. Each subscriber's communication/connection with any by other as in other discrete-address systems, is realized without intermediate commutation in common/general/total frequency channel. In system are utilized pulse noise-like signals. With such signals easily is realized the time-division multiplex, that also is realized in the system in question. The subscribers, who work simultaneously, utilize different time/temporary channels how it is provided the low level of transient interchannel interferences. The system provides the simultaneous work of 19 pairs of the subscribers (19 channels). When using a delta modulation is provided the simultaneous conduct of 10 conversations to one megahertz of the frequency band.

The discreteness of voice signal conducts by means of the taking of "cuts" with the frequency of quantization 40 kHz, i.e., through every 25 μs. Since channels 19, and segments in each channel follow through 25 μs, time, one channel, cannot be more than 25/9 = 1.3 μs. At the same time the time/temporary indeterminancy/uncertainties in system, which are composed of difference the p of the propagation time of signals from stations, which are located on different distance (signals, which go over the routes of different extent, they enter receiver displaced through time), due to pulse effect are considerably more.

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In order to avoid the difficulties, caused by these reasons, in system is utilized the method of the accumulation of cuts. The large amount of cuts is accumulated and is memorized, and then it transmits in the form of the packet, which contains the information about 400 adjacent cuts. Receiver in turn has a device, which memorizes the information about these 400 cuts and then issues alternately each cut through 25 μs.

With this method immediately transmits the information about 400 cuts and to the transmission of the cycle of the packets of 19 channels can be assigned no longer by 25 μs*, but 400 times is more, i.e., 10 ms, from which 420 μs* is utilized for the transmission of information to each channel, 100 μs* - to the shielding interval/gaps between channels and 20 μs* - to the transmission in the beginning of each cycle of the signal of synchronization in the form of word (10 μs*). The word of the signal of synchronization record/fixes the beginning of new cycle. The signal of synchronization is coded with the high degree of excess for the purpose of its protection from the possible interferences from enemy. The shielding interval/gaps between the time/temporary cuts, abstract/removed to different channels (100 μs*), are selected taking into account difference in time of the arrival of momentum/impulse/pulses due to distinguish distances of transmitters (transit time of signal from one subscriber to another and vice versa with the distance between them 8 km is 54 μs*), the effects of the effect of erosion of momentum/impulse/pulse and arrival of several ray/beams, and also the need for operational provisions with errors in the time/temporary circuits of diagram.

The diagram of the transmission of the cycle of all 19 channels is represented in Fig. 5.4.3. Incidentally let us let us point out that from the viewpoint of simplicity of the solution of the problem of the accumulation of the large number of cuts in memory system the delta modulation (just as pulse-code modulation) has the definite advantages, since with these forms of modulation there is no need for storage of information for the relation to pulse amplitude.

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In the communicating system in question are possible the following fundamental mode/conditions of the use of time, abstract/removed to each channel during the cycle:

- over channel occurs the exchange of information; in this case during time/temporary cut 420 μs transmits first the 15-marking group of address pulses, which determines the necessary subscriber, and then the group of the delta pulses, carrying information;

- on channel exchange by information does not go (pause), but

channel by this pair of the subscribers is occupied; in this case after address group follow the narrow pulses, which fix the employment of channel, i.e., not giving possibility to other subscribers to engage the channel, which in the absence of the momentum/impulse/pulses of employment could be accepted as free;

- along channel transmits the signal of interruption, realized by introduction into the usual call of supplementary redundancy; for example, if usually call begins from the repetition of the 15-pulse address of five times in a row, then the command/crew of interruption can consist of the repetition of the address of 10 times.

In considered communicating system both working between themselves correspondents utilize one and the same time/temporary channel, i.e., actually is conducted simplex communication; however in this case is retained the possibility of interruption. If necessary for the conduct of duplex communication it is necessary to have the dual assembly of instrumentation and to occupy two time/temporary channels.

Let us examine now briefly, how takes place the work of the

system in question. For a call from one or another subscriber is compose/collected called subscriber's address. Special device - "selector", trying alternately channels, finds free channel. Along this channel transmits the line signal.

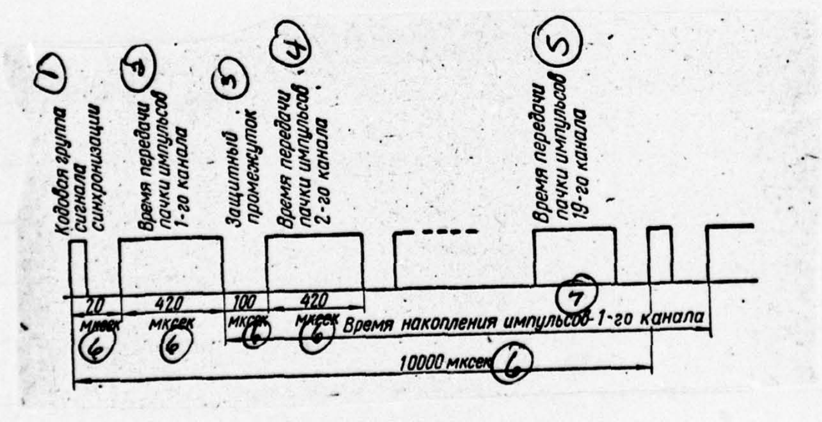

Fig. 5.4.3.

Fig. 5.4.3.

Key: (1). The word of the signal of synchronization. (2). Duration of transmission of the burst of pulses of the 1st channel. (3). Shielding interval/gap. (4). Duration of transmission of the burst of pulses of the 2nd channel. (5). Duration of transmission of the burst of pulses of 19-go channel. (6). μs. (7). Storage time of the momentum/impulse/pulses of the 1st channel. Page 246.

Thus far subscriber will not answer, in each cycle is repeated the line signal - address. Address occupies the only part of the time/temporary channel; in it remains the time ("place") for called subscriber's answer/response. Immediately after removal/taking the tube with the called subscriber line signals cease and over channel go the momentum/impulse/pulses, which indicate the employment of channel, whereupon channel is ready for the exchange of information.

Asynchronous sampled-data system. Figure 5.4.4 gives the simplified block diagram of the receiveting-transmit device of movable VHF radio station, using operation by pulse noise-like signals in the common/general/total channel of frequencies, the data on which are published into 1965 [19].

In the schematic of station is utilized pulse-phase modulation and isolation/liberation of signals with the aid of the discrete matched filters. The use of pulse noise-like signals, as is known, in a number of cases turns out to be more advisable in comparison with the use of continuous noise-like signals. Is facilitated the joint operation of several stations, especially the stations of large and small power, since the mixing momentum/impulse/pulses of large power can be between low-power. The analysis and the simulation of systems of such type show that in 13 calls in band 1 MHz is provided the signal-to-noise ratio at the output/yield used device of approximately 13 dB [19].

Let us turn now to the explanation of the work of block diagram. During the first stage the transmitted voice signal is converted into the momentum/impulse/pulses, following with frequency 8 kHz, i.e., with the interval 125 μs, during which is emitted noise-like signal, the carrying information about this momentum/impulse/pulse. Noise-like signal is form/shaped from the sequence of 63 bits (subpulses) with duration on 0.4 μs each, form/shaped by 17-stepped register. Switching the step/stages, from which is realized feedback (transfer of momentum/impulse/pulses on the input of register), i.e., the selection of address, is realized by an index register. 63-marking impulse duration 0.4 X 63 = 25.2 μs. This

momentum/impulse/pulse undergoes modulation on position on the axis of time (pulse position modulation). The maximum deviation for the pulse duration 25.2 μs and the distances between them 125 μs cannot exceed ± 49.9 μs. Page 247.

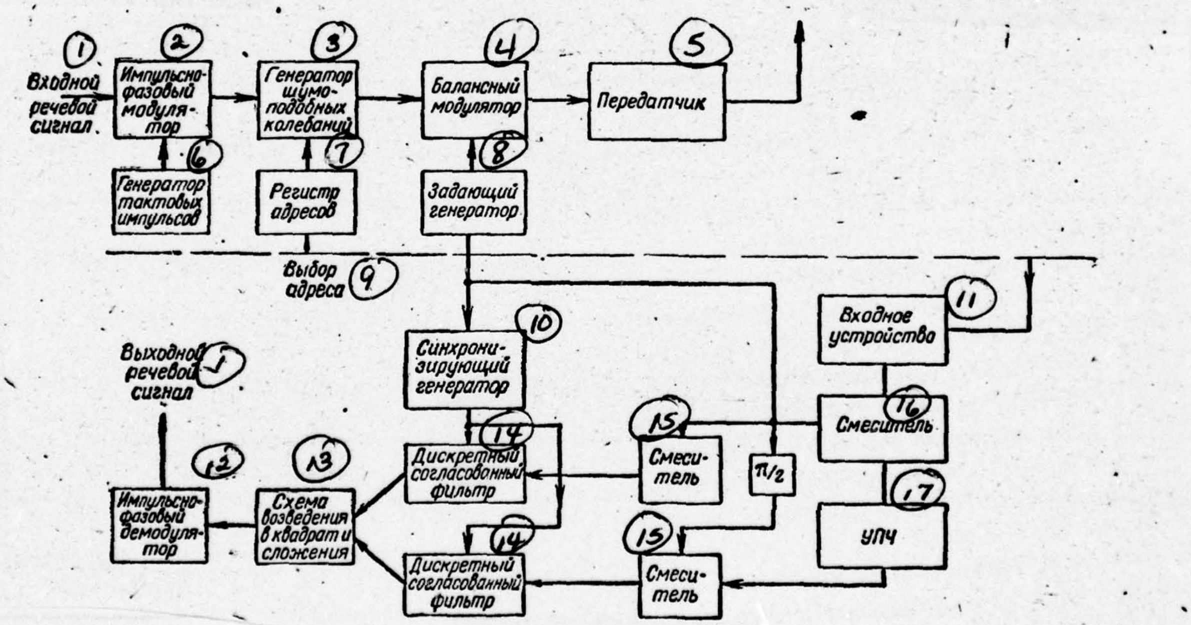

Fig. 5.4.4.

Fig. 5.4.4.

Key: (1). Input voice signal. (2). Pulse-phase modulator. (3). Generator of noise-like oscillations. (4). Balanced modulator. (5). Transmitter. (6). Clock pulse generator. (7). Index register. (8). Master oscillator. (9). Selection of address. (10). Synchronizing generator. (11). Input device. (12). Pulse-phase demodulator. (13). Diagram of squaring and additions. (14). Discrete matched filter. (15). Mixer. (16). Mixer. (17). UPCh. Page 248.

The noise-like and modulated momentum/impulse/pulses enter the balanced modulator, to which will be feed/conducted also the voltage of carrier frequency 280 MHz. Noise-like pulse signals are amplified in the output stage of transmitterand are emitted by its antenna. The sequence of 63 subpulses completed in transmitter, corresponds to the discrete code of the tuning of matched filter the receiver of that correspondent with whom is conducted the communication/connection. For the remaining receivers whose filters are not inclined to this code, this signal it is simply noise. In receptor, as is evident from block diagram, the taken signal is converted first into signal 31 MHz frequency, and then it is detected by the diagram, which is two mixer in which is realized quadrature detection. The latter removes the effect of signal fading at the output/yield of receptor. The isolated

subpulse are supplied to the matched filters, carried out on the basis of shift registers to 63 discharges. In these filters occurs folding subpulses into one momentum/impulse/pulse by duration 0.4 μs. These momentum/impulse/pulses from the output/yields of discrete filters enter the diagram of squaring and additions, whereupon to pulse-phase demodulator.

Let us note that the advantages of the discrete matched filters include simplicity of their realization in comparison with the usual analog matched filter. This is achieved by their construction as it was shown above, on the base of the widely propagated cell/elements of electronic computers. At the same time of the characteristic of the discrete matched filters with respect to the provision for a maximum of the ratio of the power of signal to noise at the torque/moment of reading very close to the appropriate characteristics of analog filters. At the same time is an essential difference between the properties of discrete and analog matched filters. Most essential of them - the discrete filters do not provide effective work under conditions of multiple-beam characteristics.

The more detailed information about the principles of construction, the characteristics of the discrete matched filters and

the results of their comparative analysis with analog filters can be found in works [19, 54].

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System of broadband communication/connection with pulse-frequency modulation. In this system [46] as and in others the type in question, also is utilized common frequency channel and possibly the establishment of the communication/connection of each with each. The system has to 1500 address codes and provides simultaneous conversation to 50 subscribers. The transmission of information is realized by the momentum/impulse/pulses, the frequency of high-frequency filling of which is determined by the modulating voltage.

The simplified block diagram of the transmitting part is given in Fig. 5.4.5. Voice signal in this diagram will be feed/conducted to four gates, which alternately are open/disclosed by the momentum/impulse/pulses, which enter from clock pulse generator. At the output/yield of gates are obtained the momentum/impulse/pulses, in their amplitude proportional to the values modulating of

cascade/stages. These momentum/impulse/pulses are record/fixed in memory units and enter the second group strobing devices that are open/disclosed by the momentum/impulse/pulses of address code in the removal/outlets of delay line, to course of which will be feed/conducted each fifth momentum/impulse/pulse from cadence generator. From the output/yield of these gates of voltage they will be feed/conducted to the control system (frequency shift key), which affects the frequency of the assigning generator of transmitter. As a result the frequency of the master oscillator at the torque/moments of the admission of the strobe pulses, following according to address law, turns out to be proportional are instantaneous to the values of the modulating voltage. The word of momentum/impulse/pulses from delay line strobes also the voltage, which enters from the master oscillator the power amplifier. At output/yield the mind is obtained address group of pulses, whereupon each address momentum/impulse/pulse is emitted at the frequency, determined by the instantaneous value of the modulating voltage at the periodically being repeated moments of time.

In receptor (Fig. 5.4.6) the taken signals enter simultaneously the amplitude and FM discriminators. After amplitude detection address momentum/impulse/pulses will be feed/conducted to delay line whose removal/outlets are included mirror with respect to delay line,

which develops address group in the diagram of the transmitting part.

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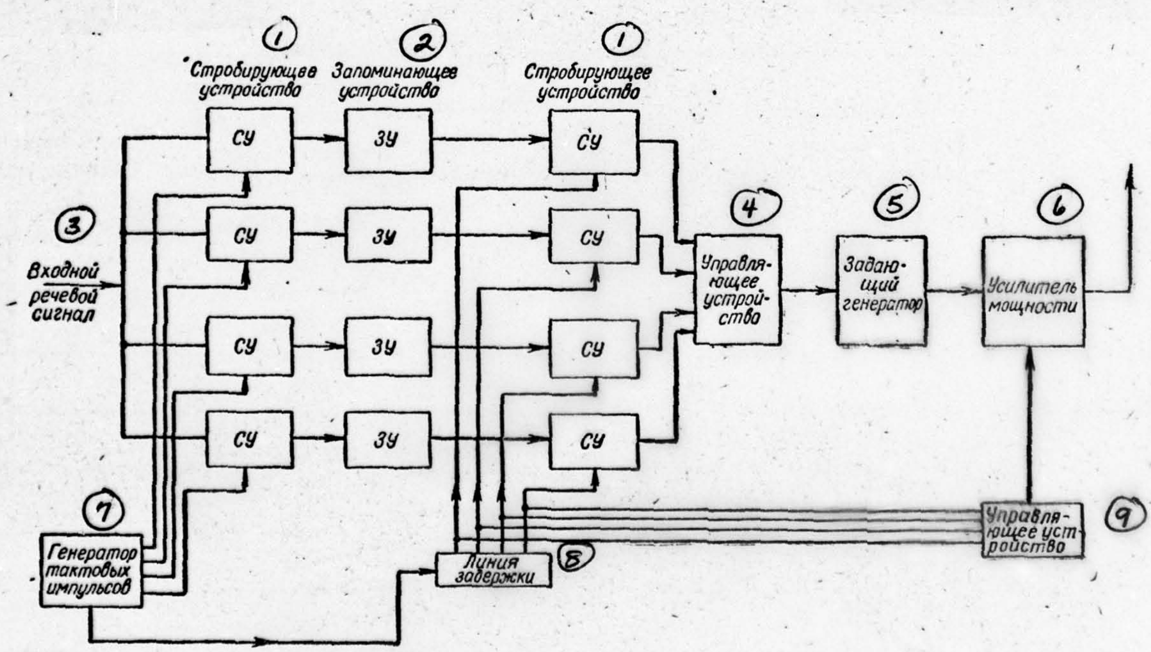

Fig. 5.45.

Fig. 5.4.5.

Key: (1). Strobing device. (2). Memory unit. (3). Input voice signal. (4). Control system. (5). Master oscillator. (6). Power amplifier. (7). Generator of cadence momentum/impulse/pulses. (8). Delay line. (9). Control system.

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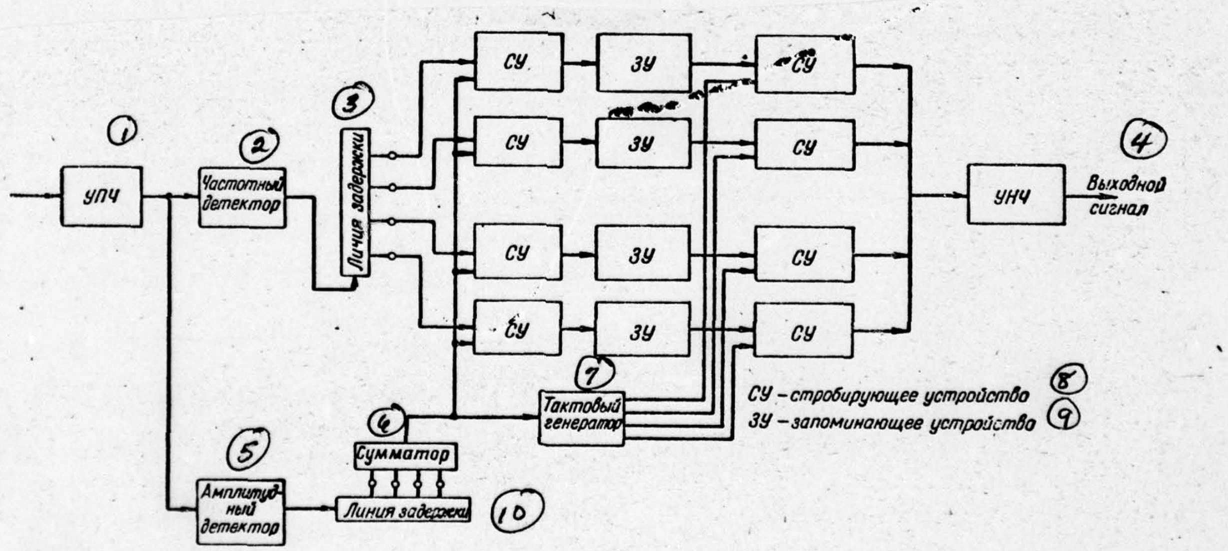

Fig. 5.4.6.

Fig. 5.4.6.

Key: (1). UPCh. (2). FM discriminator. (3). Delay line. (4). Output signal. (5). Amplitude detector. (6). Summator. (7). Cadence generator. (8). Docking system - strobing device. (9). ZU [3У ~~ЗУ~~ - memory unit] - storage device. (10). Delay line./ Page 252.

At the output/yield of this delay line is obtained resultant momentum/impulse/pulse only in the case of the arrival of correct address group from subscriber. This resulting momentum/impulse/pulse is supplied to the group of gates, to which will be feed/conducted the momentum/impulse/pulses from the second delay lines, to input of which will be feed/conducted the modulated in pulse amplitude, obtained at the output/yield of FM discriminator. The modulated along pulse amplitude are passed through gates to the memory units, when they coincide with resulting momentum/impulse/pulse with the output/yield of the first delay line. Is obtained the sequence of the modulated in pulse amplitude. These momentum/impulse/pulses are record/fixed on memory units. The gates of the second group are open/disclosed consecutively at intervals 100 μs and, therefore, to L.F. to the amplifier through every 100 μs enter the modulated through pulse amplitude. The consecutive opening of the gates of the second group is provided by the cadence generator, synchronized by

the resulting momentum/impulse/pulse from the first delay line. The utilized in system momentum/impulse/pulses have duration of 1 μs.

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Pages 253-278.

Chapter 6.

COMMUNICATION

COMPARATIVE EVALUATION OF DIFFERENT BROADBAND RADIO[redacted] SYSTEMS.

§6.1. The potential interference rejection of different broadband systems.

In this paragraph is carried out comparative analysis of potential possibilities of the different types cf broadband systems. Since all known broadband systems are intended for the transmission of discrete or continuous report/communications with the preliminary application/use of one method or the other of discreteness, are examined the only transmission systems of discrete information with the use of a binary system of signals. The primary attention is given to the case of the effect of the additive fluctuating interference, approximated by normal white noise, in channels with the constant parameters.

It is noted [for 36] that for such channels optimum is the mutually correlated system of coherent reception. Since in communicating systems the account of the initial phase of received signal in a number of cases is difficult, considerably simpler in realization are the mutually correlated systems of incoherent reception. The freedom from interference of the indicated mutually correlated systems under conditions of single-ray and multiple-pronged channels was examined into §3.4. Figure 6.1.1 shows the constructed according to formula (3.4.13) dependence of the

probability of error p in the optimum system of incoherent reception on h^2 ratio of the energy of the cell/element of signal to the spectral density of additive fluctuating interference at the input of receiver in single-ray channel (n = 1) with the constant parameters when using signals with active pause (curve 1). In channels with multiple-beam characteristic $(n > 2)$ the freedom from interference of such systems increases proportional to an increase in the energy of useful signal because of the accumulation of the incoming ray/beams. For channels with the multiple-pronged propagation of close to optimum is counted the system of the type "Rake".

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Let us compare with the optimum mutually correlated systems the freedom from interference of discrete-address (asinxronno-address) and autocorrelation broadband systems.

Discrete-address systems are intended for the realization of the communication/connection of the large number of subscribers in the overall frequency band. They are nonoptimal mutually correlated

systems, since for a decrease in charging range in them are utilized binary signals with passive pause, i.e., each subscriber's transmitter during the transmission of information can emit the broadband signal only of one determined form in alternation with pauses (absence of emission/radiation).

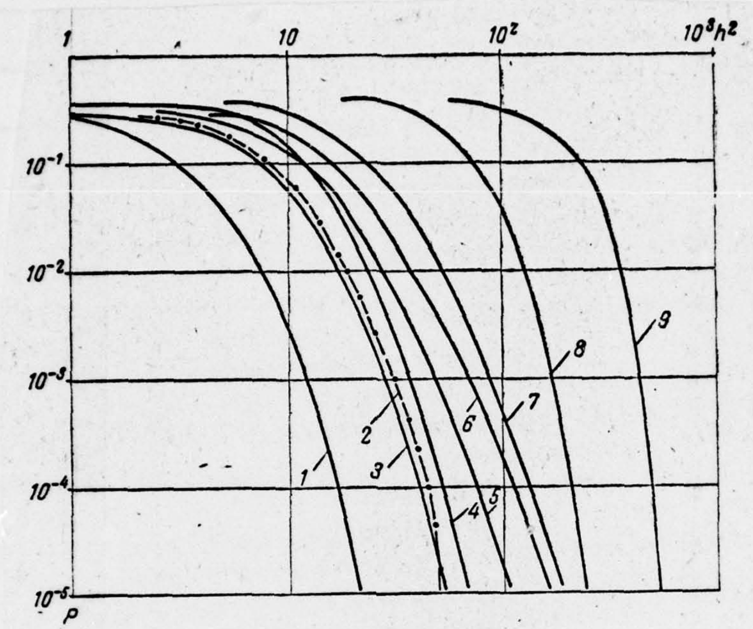

Fig. 6.1.1.

[redacted] Page 255.

By the best freedom from interference of all discrete-address systems in channel with the constant parameters and the fluctuating interference possess, apparently, systems with the noise-like phase-code-pulsed signals and the discrete matched filters, and also systems with separate processing the orthogonal components of received signal and the quadratic addition of the voltages of the separate branches of processing.

Figure 6.1.1 shows the dependence of the probability of error on value h^2 for a system with the discrete matched filters (curve 2), constructed according to the experimental data [19]. The curves 3 and 4 characterize the freedom from interference of system with separate processing and the quadratic addition of the output voltages in the case respectively of three and five branches of processing (see Fig. 5.3.2).

The worst freedom from interference possess, apparently, address systems with separate processing the orthogonal components of received signal, in which the solution to the transmission of signal

is received only with excess as output potential of each branch of the determined threshold level (see Fig. 5.3.3). In Fig. 6.1.1 the curves 5 and 6 characterize the freedom from interference of such systems for the cases respectively of three and five branches of processing. Let us note that the last/latter method of reception is realized in system RACEP. Therefore curved 5 and 6 Fig. 6.1.1 characterize the potential interference rejection of this system in channel with the constant parameters and the fluctuating interference in the transmission of discrete information by the alternation of premise/impulses and pauses equal to duration.

Finally, among broadband systems large group they compose autocorrelation systems. In these systems for the formation of the transmitted signal can be used the noise in the assigned frequency band, which substantially simplifies their realization. As an example of the realization of broadband autocorrelation system can serve examine into §4.2 diverse variants of systems with correlation-time modulation. The potential interference rejection of such systems was analyzed into §4.3. Figure 6.1.1 gives probability curves of error p from h^2, designed by formula (4.3.18) for cases $FT = 10^2$ (curve 7), $FT = 10^3$ (curve 8) and $FT = 10^4$ (curve 9).

Page 256. Let us note that formula (4.3.18) corresponds to use in the autocorrelation system of opposite signals, and therefore it determines the maximally attainable freedom from interference of these communicating systems.

Comparative analysis of curves of Fig. 6.1.1 attests to the fact that in channels with the constant parameters and the fluctuating interferences the best freedom from interference possess the mutually correlated systems, worst - autocorrelation. Discrete-address systems on freedom from interference occupy the intermediate position between the optimum mutually correlated and autocorrelation systems. The energy loss of discrete-address systems in comparison with optimum mutually correlated with the probability of error $p \geqslant 10^{-4}$ composes depending on the construction of diagram 4-8.8 dB. For autocorrelation systems this value will be not less than 9-15 dB.

§6.2. On the work of broadband communicating systems in the loaded frequency band.

The freedom from interference of binary communicating systems in single-ray channel under the influence only of normal fluctuating

interferences does not depend on the band of frequencies F of those utilized for the transmission of the information of signals. It is determined in this case only by ratio of the energy of received signals to the spectral density of the fluctuating interference

$$h_0^2 = \frac{P_0 T}{\nu^2}. \tag{6.2.1}$$

Recall that in two limiting cases of the state of channel in the absence of fadings and the presence of Rayleigh fadings the probability of the error of piece-by-piece reception in mutually correlated systems with active pause is equal to respectively

$$\left.\begin{aligned} p &= \frac{1}{2} e^{-\frac{h_0^2}{2}}; \\ p &= \frac{1}{h_0^2 + 2}. \end{aligned}\right\} \tag{6.2.2}$$

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However, in multiple-pronged channel the correct selection of the bandwidth of signals F becomes essential. The necessary preliminary condition of the separation of the incoming ray/beams and accumulation of their energy is the selection of this band F in order that it would satisfy condition (see §3.3)

$$F > \frac{1}{\Delta t_{min}}, \qquad (6.2.3)$$

where Δt_{min} - the minimum time lag between separate ray/beams.

The real communication channels are always subjected to effect not only fluctuating, but also the large number of the interferences of radio aids - the concentrated interferences. It would seem that the expansion of the band of frequencies of the utilized signals must bring in such channels to a decrease in the freedom from interference of communicating systems. However, more in-depth analysis, as this partially has already been discussed in §3.4, shows that in the loaded ranges the broadband communicating systems can ensure the better/best freedom from interference, than narrow-band.

In this paragraph, following the results [4, 42], is carried out comparative analysis of freedom from interference and reliability of the operation of broadband and narrow-band systems in the loaded

frequency bands, and also is given comparative evaluation of broadband systems with the narrow-band systems, in which operating frequencies are selected by cut-and-try method on the basis of the analysis of the state of range.

Let us examine the question concerning a comparative evaluation of the freedom from interference of broadband and narrow-band systems. The presence of the interferences of radio aids leads to the fact that the values of the spectral density of interferences at the input of receptor and, consequently, also h^2_0 are the random variables, which depend on the state of charging range and character of the concentrated interferences. It is natural that i the probabilities of errors (6.2.2) in this case also will be random variables. For the search of the unconditional probabilities of the errors in this situation, we utilize the following prerequisite/premises.

1. Charging kV of range is so/such great, that in all operating range, selected for the transmission of information, there will not be the vacant places, immune to to the effect of the concentrated interferences.

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In other words, the spectral density of such interferences is a continuous function of frequency. This condition describes the heaviest situation from the viewpoint of the work of broadband systems.

2. The rate of change in the spectral density of interferences in time considerably lower than the speed of transmission of information, i.e., the spectral density of interferences is constant during the duration of the cell/element of signal T, but is random from one cell/element to the next.

3. Since the interferences from the stations, which work at the adjacent frequencies, are independent random quantities, the correlation spectral density function of the concentrated interferences becomes zero with a difference in frequencies Δf, in the equal to bandwidth, occupied by one station. The interval Δf is the interval of the correlation of interferences of frequency. For a

skip band Δf = 1-2 kHz.

A sufficently good approximation for the normalized correlation function of interferences on frequency ρ (f), that allows us to establich the fundamental laws governing their affect on broadband and narrow-band systems, is the expression of form [4]

$$\rho(f)=\begin{cases}\left(1-\frac{|f|}{\alpha F}\right), & {}^{(1)}\text{при } |f| \leqslant \alpha F;\\ 0, & {}^{(1)}\text{при } |f| > \alpha f.\end{cases} \qquad (6.2.4)$$

(1) with

Here $\alpha = \Delta f/F$ is ratio of the interval of correlation of frequency between interferences to the bandwidth of the signals, utilized for the transmission of information.

Under the made assumptions it is possible to count that the spectral density of interferences at the input of receptor is determined by the relationship/ratio

$$\nu_{\text{п}}^2 \leqslant \nu^2 e^y, \qquad (6.2.5)$$

where y - is normal random variable with probability density

$$W(y)=\frac{1}{\sqrt{2\pi}\sigma_y}e^{-\frac{y^2}{2\sigma_y^2}}; \qquad (6.2.6)$$

i.e. having zero mathematical expectation and the dispersion, which during satisfaction of condition (6.2.4) takes the form

(1) with

$$\sigma_y^2=\begin{cases} 4\sigma^2\alpha\left(1-\frac{\alpha}{3}\right),^{(1)} \text{при } 0\leqslant\alpha\leqslant 1; \\ 4\sigma^2\left(1-\frac{1}{3\alpha}\right),^{(1)} \text{при } 1<\alpha<\infty. \end{cases} \qquad (6.2.7)$$

Page 259. Here σ^2 is dispersion of a change in the jamming intensity in the frequency of the communication channel. Then, substituting in (6.2.1) value (6.2.5), we obtain that the random variable of the ratio of the energy of signal to the spectral density of the affecting interferences is equal to

$$h^2 \geqslant h_0^2 e^{-y}. \qquad (6.2.8)$$

Let us note that the probabilities of errors in the communicating systems in question are the monotonically decreasing functions of value h^2. Therefore subsequently let us examine values h^2, which correspond to equality in (6.2.8), that determines the worst freedom from interference and is upper limit for the probability of error at other values h^2, greater than right side (6.2.8). The further unconditional probability of error can be found now by substitution in relationship/ratios (6.2.2) of value h^2 from (6.2.8) instead of $h^2{}_0$ and by the averaging of these relationship/ratios in terms of all possible values of quantity h^2 with the aid of random number distribution y.

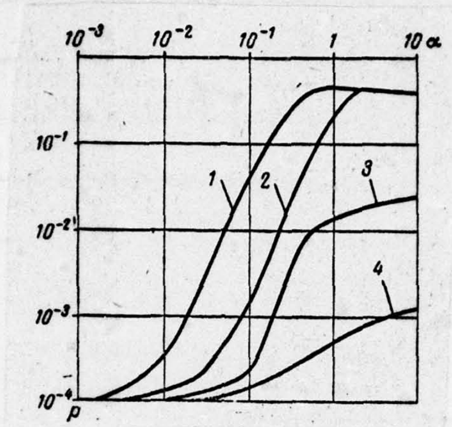

Fig. 6.2.1.

[redacted] Page 260.

Then

$$
\left.\begin{aligned}
p &= \frac{1}{2}\int_{-\infty}^{\infty} e^{-\frac{h_0^2 e^{-y}}{2}} W(y)\,dy;\\
p &= \int_{-\infty}^{\infty} \frac{1}{h_0^2 e^{-y}+2} W(y)\,dy,
\end{aligned}\right\} \qquad (6.2.9)
$$

where the probability density W (y) is determined by formula (6.2.6). These probabilities of the errors were calculated in work [4]. Unfortunately, they in elementary functions are not expressed. The probabilities of error by formulas (6.2.9) can be calculated with the aid of ETsVM [ЭЦВМ [redacted] - digital computer] or tables [41].

Analysis (6.2.9) shows that the probability of the error in this

case depends not only on h^2_0, but also on σ^2_y, by determined formula (6.2.7). In other words, in the case of the effect of the concentrated interferences the probability of the error is function not only h^2_0, but also it depends on α - the ratio of the interval of correlation of frequency between interferences to bandwidth F of the utilized signals. Figure 6.2.1 shows the dependences of the probability of error (6.2.9) for the nonfading signal on value α for $\sigma^2 = 4$ (curve 1) and $\sigma^2 = 1$ (curve 2). Value h^2_0 in this case is fix/recorded and equally h^2_0 - 17, which corresponds to the maximum probability of error 10^{-4} in channel only with fluctuating interferences. As can be seen from figure, deterioration in the freedom from interference with respect to maximum can be very considerable with insufficiently to large band of signals ($\alpha > 0.01$). The freedom from interference of reception very close to maximum, if value F exceeds the bandwidth of jamming stations Δf not less than 100 times ($\alpha \leqslant 0.01$). If Δf = 1-2 kHz, then for the transmission of information in the loaded frequency band one should utilize broadband signals with band 100-200 kHz. Such signals make it possible to divide and to accumulate ray/beams with relative time lag 5-10 μss.

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Figure 6.2.1 according to the results [4] gives also dependence curves of the probability of the error with Rayleigh signal fading as a function of α for $\sigma^2 = 4$ (curve 3), $\sigma^2 = 1$ (curve 4) and value $h^2_0 = 10^4$ which corresponds to the maximum probability of error 10^{-4} in channel only with fluctuating interference. From these curves it is evident that with Rayleigh signal fading the selection of bandwidth F not as strongly affects freedom from interference in comparison with the preceding/previous case. The selection of value F 10 times greater the width of the spectrum of jamming stations ($\alpha = 0.1$) is sufficient for providing a probability of the error, close to its limiting value in channel only with fluctuating interference.

As a whole Fig. 6.2.1 shows that when using in the loaded range of broadband signals with sufficiently large band F ($\alpha \ll 1$) is provided the considerably higher freedom from interference, than when using narrow-band signals ($\alpha = 1$).

Let us examine further the question concerning the reliability of the operation of broadband and narrow-band communicating systems in the loaded frequency band. Under the reliability of operation in this case is understood probability that the quality of communication/connection (probability of erroneous reception) will be

better determined preestablished value. Above it was shown, that in the loaded frequency band the probability of the error in the systems in question is the random variable, which according to (6.2.2) and (6.2.8) is equal to

$$p=\frac{1}{2}\exp\left[-\frac{h_0^2}{2}e^{-y}\right] \qquad (6.2.10)$$

- fo[illegible] fading and

$$p=\frac{1}{h_0^2 e^{-y}+2} \qquad (6.2.11)$$

- for the fading according to Rayleigh law signal.

In these expressions y - the normal random variable, determined on (6.2.6).

During the transmission of information the communicating system must possess the determined reliability - the probability of error p in it must be less than certain permissible value p_{om}:

$$p \leqslant p_{\text{om}}. \qquad (6.2.12)$$

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With the nonfading signal to expression (6.2.12) is equivalent, as this follows from (6.2.10), the inequality

$$\frac{1}{2} \exp\left[-\frac{h_0^2}{2} e^{-y}\right] \leqslant p_{\text{om}}. \qquad (6.2.13)$$

Whence, twice logarithmizing the right and left sides, we find

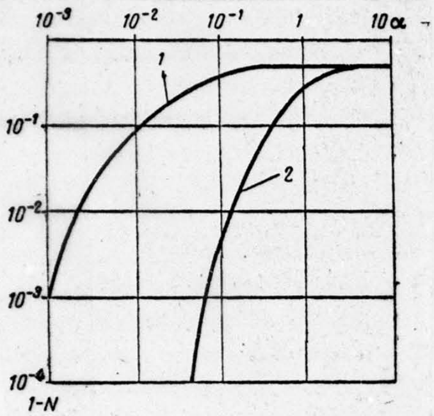
$$y \leqslant \ln \frac{h_0^2}{2} - \ln \ln \frac{1}{2p_{\text{ош}}} = A. \qquad (6.2.14)$$

The probability of the fulfillment of this inequality is reliability of communication/connection N. Consequently,

$$N = \int_{-\infty}^{A} W(y)\,dy = \frac{1}{2}\left[1 + \Phi\left(\frac{A}{\sigma_y}\right)\right]. \qquad (6.2.15)$$

Figure 6.2.2 shows the dependence (curve 1) of reliability N on value α - the ratio of the interval of correlation of frequency between interferences (bandwidth of jamming stations) to the bandwidth of **the utilized signals.** This curve is constructed for $h^2_0 = 17$, $\sigma^2 = 1$ and $p_{\text{ош}} = \frac{5}{2} e^{\frac{-h_0^2}{2}}$.

Fig. 6.2.2.



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From the figure one can see that for providing sufficiently high reliability N ⩾ 0.9 transmitted signals must have a frequency band, 100 times exceeding band of frequencies of jamming stations (α = 0.01).

In the case of the fading signal of error probability p than less assigned $p_{\text{он}}$, if is fulfilled the inequality

$$\frac{1}{h_0^2 e^{-y}+2} \leqslant p_{\text{ош}}. \qquad (6.2.16)$$

Whence for the reliability of communication/connection N with $p_{\text{ош}} \ll 1$.

$$N \approx \frac{1}{2}\left[1+\Phi\left(\frac{\ln h_0^2\, p_{\text{ош}}}{\sigma_y}\right)\right]. \qquad (6.2.17)$$

Figure 6.2.2 (curve 2) shows dependence of N on α for the fading signal with $h^2{}_0 = 10^3$, $\sigma^2 = 1$, $p_{\text{ош}}=\frac{5}{h_0^2+2}$. From the figure one can see that in this case the high reliability (more than 0.9) is provided with $\alpha \leqslant 0.1$, i.e., with the band of frequencies of the transmitted signals, 10 times by that exceeding the band of frequencies of jamming stations.

Thus, the use of broadband signals makes it possible to ensure the higher reliability of communication/connection in the handled

range in comparison with narrow-band signals.

At the same time when using narrow-band signals it is possible also to attain as small as desired probability of erroneous reception, if we select the operating frequency of narrow-band station on least impregnated by interferences range frequencies. this selection of operating frequency is based on selection with its cut-and-try method on the basis of the analysis of the state of range. Let us show that i in this case application/use of broadband signals gives the determined advantages. Let us determine the number of sample/tests, necessary for provision in narrow-band system with the assigned quality of the communication/connection of the required reliability.

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Communication/connection will be realized with the probability of the error, which does not exceed p_{om}, if at least on one of m of the selected frequencies the jamming intensity it is have sufficiently low value. the probability this

$$N=1-\left[\int\limits_{p_{\text{ош}}}^{\infty} p(y)\,dy\right]^m, \quad (6.2.18)$$

where p (y) is determined by formula (6.2.10) with that which not fade and by formula (6.2.11) with the fading signal.

Figure 6.2.3 depicts to the dependence of the necessary number of attempts m on the bandwidth of signals (parameter α) for N = 0.999, $p_{\text{ош}}=5p(h_0^2)$, σ^2 = 1 and $h^2{}_0$ = 17 (curve 1), and also $h^2{}_0 = 10^3$ (curve 2). From the figure one can see that for the nonfading signal (curve 1) the necessary number of attempts is equal to nine with the narrow-band signal (α ~~ 1) even of five with broadband signal ($\alpha \leqslant$ 0.01). For the fading signal (curve 2) the necessary number of attempts is respectively equal to six and less than two (with $\alpha \leqslant$ 0.1).

Thus, broadband system makes it possible to obtain the assigned reliability of communication/connection with the substantially

smaller number of sample/tests for the selection of operating frequency. To this one should add that the narrow-band system with the selection of operating frequency requires, furthermore, the presence of the return duct of communication/connection of receiver to transmitter for the transmission of the information about the possibility of work on the selected frequency. Broadband communicating system in principle this channel does not require.

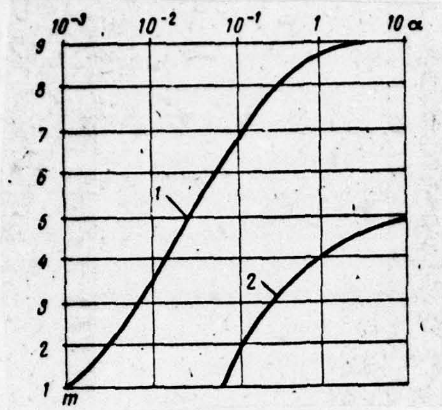

Fig. 6.2.3.

[redacted]. Page 265.

Along with examined higher very large charging range in practice are possible also the cases, when the receptor affects the relatively small number (one - heel) of the concentrated interferences. Furthermore, during large charging range to receptor, as a rule, can affect the concentrated interferences, which are sharply isolated against the background of their totality. Such single concentrated interferences have power, considerably exceeding power of other interferences, and their spectra partially or completely they coincide with the spectrum of signal.

The effect of the single concentrated interferences on communicating system depends on their statistical properties and on the parameters $h_{\text{п}}^2/FT$, where $h_{\text{п}}^2$ is ratio of the energy of interference to the spectral density of fluctuating noise, FT is a base of system. The interference of such interferences begins to manifest itself with $h_{\text{п}}^2 > FT$ (for example, see §3.4). Hence it follows that in the broadband communicating systems, which have FT order of several hundreds or even thousand, is realized the considerably more powerful suppression of the single interferences, than in narrow-band, for which $FT \ll 10$. At the same time the powerful concentrated interferences with $h_{\text{п}}^2 > FT$

can cause a considerable decrease in the freedom from interference, also, in broadband systems.

The most radical means for combat with such interferences is, apparently, the rejection at the input of the receiver of the sections of the band of signal F, subjected to the effect of the concentrated interferences. In this case, if rejected sections are sufficiently narrow, there will not be a noticeable decrease in the energy of the adopted broadband signals. Rejection can be sufficiently effectively used in mutually- and autocorrelation systems. Unfortunately, it is not applicable in discrete-address systems on the strength of the special feature/peculiarities of formation and processing in them signals (see §5.2). For these systems, apparently, it is most expedient to utilize the various kinds of the diagram of the compensation for interferences, that are the component part of the decisive receiver circuit.

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§6.3. On the reticence of the broadband transmissions in the real communication channels.

On the basis of the given in the preceding/previous chapters data on freedom from interference and special feature/peculiarities of the practical realization of the different types of broadband systems it is possible to make some conclusions about the possibilities of the concealed work of broadband communicating systems and first of all on the possibilities of providing a reticence of the very fact of the work of broadband radio link.

As it was noted in §1.5, in broadband communicating systems the reception of discrete information can be realized at the power of signal smaller than the power of fluctuating interferences in the utilized frequency band. In this case the reticence of broadband transmission quantitatively can be described at the assigned authenticity of reception and speed of transmission of information by the value of the ratio of the power of signal P_c to the power of fluctuating interferences $P_п$ in the utilized band of frequencies, i.e., as this follows from formula (1.5.2), by value

$$q_0 = \frac{P_c}{P_\text{п}} = \frac{h^2_\text{тр}}{FT}, \qquad (6.3.1)$$

where FT - the base of system; $h^2_\text{тр}$ - the required for obtaining the assigned probability of error ratio of the energy of signal to the spectral density of fluctuating interference.

Under the assigned probability of the error in this paragraph, let us consider the probability of error equal to 10^{-4}-10^{-5}. As can be seen from relationship/ratio (6.3.1), the relation $P_c/P_\text{п}$ unambiguously is determined by value $h^2_\text{тр}$ and by the value of the base of system. Moreover for providing a reticence of broadband transmission it is necessary to satisfy the condition

$$q_0 < 1. \qquad (6.3.2)$$

In this case the signal is arrange/located below the level of fluctuating noises, as this is schematically shown in Fig. 6.3.1.

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From the examined broadband systems the greatest freedom from interference possess mutually correlated systems.

In channels with the constant parameters (see §3.4) and only fluctuating interferences for obtaining the probabilities of errors p = 10^{-4}-10^{-5} value h^2_{TP} independent of the character of the construction of the decisive receiver circuit does not exceed 20-30. It is considered that such systems possess the greatest potential possibilities from the viewpoint of the reticence of the work of broadband radio link. Under these systems condition (6.3.2) is satisfied already at values FT order of several dozens.

In autocorrelation systems depending on the diagram of processing signal in the receptor of value h^2_{TP} they compose the value of the order of several hundreds or even thousand. Therefore for the

concealment of the fact of broadband transmission such systems require incomparably the large values of the base of system. Substantially important is that fact that both the mutually correlated and autocorrelation systems can in principle ensure the concealment of the fact of transmission. Discrete-address systems are deprived of this possibility. On the strength of the specific character of formation and decoding of signals the transmission of information in discrete-address systems is realized only with $q_0 \gg 1$. Therefore in connection with such systems it is possible to speak only about the reticence of the fact of the presence of information in signal (see §1.5) which can be reached with signal conditioning by the selection of the corresponding methods of coding in frequency-time range. However, in spite of the indicated deficiency/lack, discrete-address systems possess a series of remarkable properties (see Chapter 5), which make them with very convenient for a military radio communication.

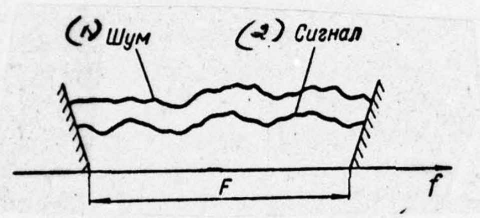

Fig. 6.3.1.

Key: (1). Noise. (2). Signal. Page 268.

The concealment of the fact of the presence of the broadband transmission in the real channels of communication/connection, i.e., when signal fading are present, and under the influence on the receptor not only of fluctuating, but also concentrated interferences, is largely hinder/hampered and its provision requires special measures during the construction of the information-carrying

system.

First of all let us note that the presence of signal fading lowers the freedom from interference of broadband systems, increasing thereby values h^2_{TP} and making satisfaction of condition worse (6.3.2). For example, even in mutually correlated systems during processing only one of the incoming and fading according to Rayleigh law ray/beams value h^2_{TP} it composes, as this follows from formulas (3.4.2) and (3.4.3), value $(0.25-1) \cdot (10^4-10^5)$.

The technically realizable values of base by broadband systems do not exceed, apparently, several thousands. Therefore the provision for a reticence of broadband system in an example in question is very problematic.

At the same time it should be pointed out that the accumulation of several incoming ray/beams makes it possible substantially to improve reticence. Actually, already with the accumulation of two subjected to Rayleigh fadings ray/beams the necessary values h^2_{TP} depending on the method of processing signal (see formulas (3.4.15) -3.4.17) compose values $(0.85-5.4) \cdot 10^2$. An increase in the number of

workable ray/beams even more descends $h_{тр}^2$, approaching it the appropriate values in channel with the constant parameters. The possibility of concealed transmission under these conditions substantially is improved.

Under the effect in the channel of the communication/connection not only of fluctuating, but also concentrated interferences the reticence of broadband transmission can be largely deteriorated. Specifically, from results §3.4 it follows that the freedom from interference of mutually correlated systems under these conditions substantially depends not only from the ratio of the energy of signal to the spectral density of fluctuating noise, but also from the coefficients of the cross correlation between the utilized signals and the affecting concentrated interferences. In this case for obtaining the same correctness of the adopted report/communication as in channel only with fluctuating noises, will be required the large values of the energy of signal, which, is logical, it can lead to the disturbance/breakdown of condition (6.3.2).

In the case of the simultaneous effect of single sinusoidal and fluctuating interferences as this it is possible to show, by utilizing results [30, 53], the value of the ratio of the power of

signal to the power of the fluctuating interferences

$$q \leqslant \left(1 + \gamma_0 \frac{h_n^2}{FT}\right) q_0, \qquad (6.3.3)$$

where q_0 - it is determined by formula (6.3.1) for the case of the effect only of fluctuating interferences; γ_c - the coefficient, depending on the type of the utilized signals and special feature/peculiarities of their processing in receptor; for example, with coherent reception to the coherent addition n of the ray/beams of approximately identical intensity value γ_c does not exceed n and 2n for opposite and orthogonal signals respectively; with incoherent reception with the coherent addition of such ray/beams $\gamma_c < 2n$.

In Fig. 6.3.2 with the aid of the results [30] are constructed dependences q as a function of h_n^2 for the case of the single-ray coherent reception of opposite signals $(\gamma_c=1)$. These curves are constructed for the values of base FT = 10^2 and 10^3 and the authenticity of the adopted report/communication, determined by the probability of error 10^{-4}-10^{-5}. From curve/graphs it is evident, that even so for the broadband system FT >> 1, nevertheless at $h_n^2 \gg FT$

value q becomes more than one. The latter corresponds to that fact that of noise-like signal level in band F will be above the level of fluctuating noises (Fig. 6.3.3), and consequently, the reticence of system largely deteriorates.

Let us note that with somewhat smaller requirements for the probability of error, let us say 10^{-2} or 10^{-3}, the reticence of broadband system can be improved.

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However, with contemporary requirements for the correctness of reception scarcely whether one should increase reticence by the value of the loss of authenticity.

As can be seen from Fig. 6.3.2, the reticence of broadband system can be substantially raised by an increase in the base of signal FT. An increase in the base because of the expansion of the conditional frequency band is limited in the range of short waves by the dispersed properties of the ionosphere, since as it is noted in

§1-4, value F in these conditions was limited by the values several dozen kilohertz.

The further increase FT is possible only because of an increase in the duration of the cell/element of signal T, i.e., deceleration of the transmission of information.

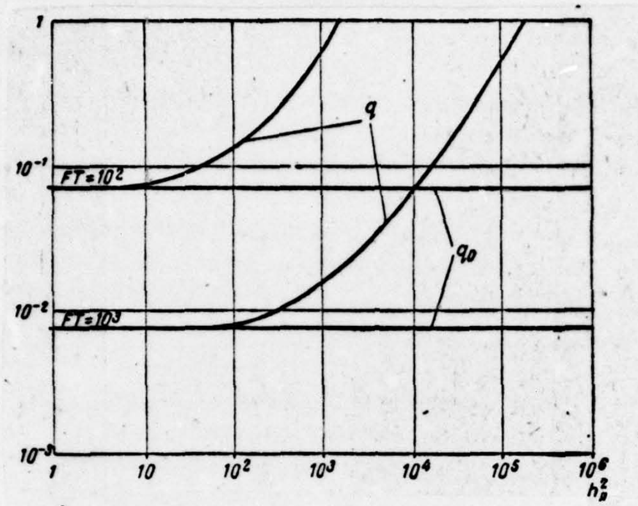

Fig. 6.3.2.

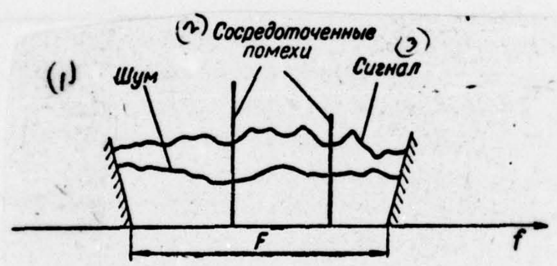

Fig. 6.3.3.

~~Fig. 6.3.2.~~

Fig. 6.3.3.

Key: (1). Noise. (2). Concentrated interferences. (3). Signal. Page 271. The curves of Fig. 6.3.2 can, in particular, serve as the illustration of the reticence of broadband system with speed of telegraphing of 50 bauds (T = 20 ms) and values F = 5 kHz ($FT = 10^2$), 50 kHz ($FT = 10^3$) or at the constant band F = 10 kHz and telegraph speeds 100 bauds (T = 10 ms) even 10 baud (T = 0.1 s).

The indicated in this example laws are valid and in the general case. Of course, with processing and accumulation of several incoming ray/beams of condition for a decrease in value q there can be considerably improved.

Finally, the radical means of providing condition $q < 1$ without a noticeable decrease in authenticity and speed of transmission of information is the exception/elimination (rejection) of the concentrated interferences at the input of receptor.

Thus, the requirements for the reticence of broadband transmission, high authenticity of the transmitted report/communication and high speed of transmission of information are contradictory. This contradiction substantially grow/rises when fadings of received signals are present, and under the effect in the channel of the communication/connection of the concentrated interferences.

The provision for the necessary reticence of the fact of the transmission of information can be reached in the work of broadband systems only by the complex of actions, including by fight with the concentrated interferences, deceleration of the transmission of information, etc. In this plan/layout sufficiently effective can be the application/use in the broadband system of the channel of feedback, which controls the level of the necessary energy of signal in accordance with the state of the information circuit. Let us note also that the detection of broadband transmission can be very hinder/hampered also in the case of $q > 1$, if is a priori unknown the level of fluctuating noises in channel. Therefore in cases when the considerations of economic order, weight and overall sizes of equipment and others are not prevailing, the reticence of the

broadband transmission can be provided for by means of masking by broadband signal, carrier of information and emitted during long time τ, signal, carrying information during sufficiently short time interval $\tau' \ll \tau$.

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CONCLUSION.

The examined in the book fundamental special feature/peculiarities and the advantages of broadband radio communication have important practical value, making it possible to actively resist the difficulties, which appear as a result of multiple-pronged propagation, closeness in ether/ester also of the organized radio jammings.

At the same time it will incorrect to assert that the further perfection/improvement and the development of broadband systems are the fundamental and only development trend of radio communication.

It is necessary to keep in mind, that even so in principle the use of broadband signals is completely consistent with narrow-band, all the same question of the simultaneous use of those and other systems in one frequency range and especially within the limits of one object of communication/connection (ship, aircraft, etc.) runs into serious difficulties. These difficulties are caused first of all by the interferences, which create broadband systems by the near arranged/located receptor in an entire emission band of broadband transmitter.

The questions of the organization of communication/connection and rational distribution of frequency range in the case of the simultaneous use of those and other systems still require deep studying. Are not completely solved also the questions of the organization of the work of the grid/network of the broadband radio stations, which use principle of address radio communication in VHF range.

the major advantages of broadband communicating systems are reveal/detected most completely when using signals with large base.

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Therefore it is clear that the tendency to obtain the maximum gain from broadband systems to a considerable degree contradicts the tendency of an increase in the velocity of the exchange of information. This determines the advisability of using broadband systems when the speed of transmission of information is not decisive, but the correctness of its reception under conditions of channel with the strongly changing parameters plays an especially important role.

let us note also the fact that the broadband communicating systems can be seriously improved in the case of their supplement stable working channel of feedback. In this case appears the possibility of operational effect on the parameters of the transmitted signal (for example, the speed of transmission of information) depending on the state of channel and which is especially important, the dynamic control of radiated power. This undoubtedly will ensure an essential decrease in the interferences

and will raise possibilities in the relation to an increase in the number of simultaneously working and jamming stations in the limited section of range. However, this use of a channel of feedback is very expedient for all other radiolink systems.

The examined broadband systems in a number of cases is realized somewhat more complex than widespread narrow-band systems. One of the promising trends, which simplify their realization and which make it possible sufficiently it is simple to solve the questions of formation and decoding of signals, is the application/use widespread cell/elements, utilized in the technology of electronic computers.

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